EXPERIMENTAL PROCEDURE FOR DETERMINATION OF Q_U, Q_{EXT}, Q_L AND COUPLING (β) OF DIELECTRIC RESONATOR (DR) AND DEVELOPMENT OF DR OSCILLATOR

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Master of Technology

by

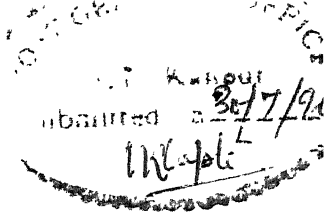
Pavan Kumar

to the

DEPARTMENT OF ELECTRICAL ENGINEERING

INDIAN INSTITUTE OF TECHNOLOGY, KANPUR

Certificate



It is certified that the work contained in the thesis entitled **EXPERIMENTAL PROCEDURE FOR DETERMINATION OF Q_U, Q_{EXT}, Q_L AND COUPLING (β) OF DIELECTRIC RESONATOR (DR) AND DEVELOPMENT OF DR OSCILLATOR**, by Pavan Kumar , has been carried out under my supervision and that this work has not been submitted elsewhere for a degree.

25 July 1996

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Abstract

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Experimental procedure to measures the coupling between DR and microstrip (β) has also been presented. An experiment has been performed using HP8410 network analyser and β, Q_{ext}, Q_L, Q_u has been measured to evaluate equivalent circuit parameters of DR. Finally, DR oscillator has been designed and fabricated.

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Acknowledgement

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Also, the inspiring lectures of Dr. Biswas and Prof. M. Sachidananda instilled in me a sense of appreciation and wonder for electromagnetics and microwaves.

I wish to thank my batchmates J. Kumar, Patni, Yash, Manish, Ashish, Jaipal, Jayesh, Sanjay, Raju, Subbu and friends Roopam, Maj. Chow, P. K. Paliwal, Maj. Sehgal and Santosh who have always stood by me during my stay in this campus.

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Chapter 1

INTRODUCTION

Microwave systems like radar, electronic warfare, navigation, surveillance and weapon guidance system are largely military in nature and research in these fields have been largely sustained by defense community throughout the globe. Microwave are also used in commercial systems like cellular phone, microwave oven, data communications and video/audio communications.

Now a days, reduction in the cost and size are main the parameters in the design consideration of microwave circuits. Although bulky waveguide and rigid coaxial lines are replaced by microstrip and striplines but former are still used in some application, such as high power transmissions line or low loss filters. In the design of microwave systems like filter, mixer and oscillator, a frequency determining element is required. Metal cavity resonator, SAW (Surface Acoustic Wave) or microstrip line can be used as frequency determining element. With the advent of high dielectric constant, temperature stable and low loss material, Dielectric Resonator(DR) can also be used for same purpose. DRs are compatible to microstrip and suspended stripline for design of microwave circuits. The size of DR is considerablely smaller $(1/\sqrt{\epsilon_r})$ than the size of empty metal cavity operating at same frequency. They are

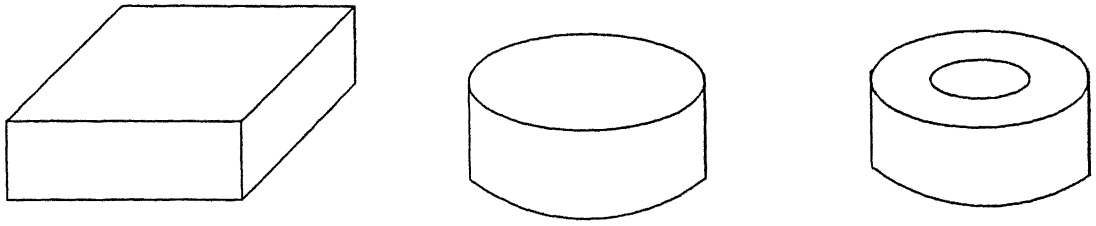


Figure 1.1: Different Shape of Dielectric Resonator

1.1 DIELECTRIC RESONATOR

R.D.Richtmayer in 1939 observed that unmetallized dielectric objects can function as dielectric resonators. Its resonant frequency, modes, and circuit properties were first verified experimentally in 1960.

DRs are made of low-loss, temperature stable, high permittivity and high Q ceramic material in a regular geometrical form. The shape of DR is usually a short and solid cylinder but one can also find tabular, spherical and paralleled as shown in figure1.1. A commonly used resonant mode in cylindrical DRs is $TE_{01\delta}$. This shape of DR is preferred over other shapes because this does not have a large number of degenerative modes and is easier to fabricate and mount in microwave circuits. Its resonant frequency is determined by its dimensions and shielding condition. Because of its small size, low price and excellent integrability in MICs, it is very useful in both active and passive circuits.

1.1.1 Quality Factor

Quality factor of a circuit is a very important parameter because it limits the overall performance of the circuit. Higher value of Q is required in design of oscillators.

The unloaded quality factor of DR placed in suspended substrate environment depends upon loss in the DRs material, conductor loss of the shielding wall and

past to compute the Q-factor [1]-[4]. Although these methods are quite accurate but are highly numerical in nature. Kajfej [5] has described method of perturbation for computing Q-factor for DR in MIC environment. This determines the Q-factor indirectly by computing the relative change in resonant frequency when the conducting walls of the DR structure are moved inwards by an amount equal to skin depth. Although this method is quite accurate but lot of caution may be required when applying this method, because these two frequencies (perturbed and unperturbed) are very close to each other.

In this thesis Q-factor is determined for generalized two layer substrate by computing the stored energy and power loss. This method is approximate but quite accurate and data calculated by this method can be used in design purposes. Mongia [6] has done same work for MIC environment, but he did not incorporated frequency dependent dielectric loss in DRs and substrates. These losses might be significant at higher frequency.

1.1.2 Temperature coefficient(T_f)

The temperature coefficient (T_f) of a DR can be controlled by modifying the composition, to be anywhere with in +9 to -9 p.p.m/^{o}c. Change in any circuit parameter can also shifts T_f by few part per million.

Table1.1 [7] shows the properties of certain material used in DR.

In table 1.1 no composition have overall superiority over the others, since many factor, such as ease of ceramic processing and ability to hold tolerances on the dielectric properties must be considered.

Material composition	Q_0	ϵ_r	Freq range (GHz.)	Temp coff. (T_f (ppm/°c))
(ZrSn)Ti oxide	10,000(4.5GHz.)	35.7-36.4	1.45-8.86	-3 to +9
BaLnTi oxide	3,000(3 GHZ.)	80.0	0.7-3.62	-3 to +9
BaZnTaTi oxide	10,000(10GHz.)	27.6-30.6	5.5-32.2	-4 to +4
Ba Titaniam oxide	6,000(4.5GHz.)	36.6-38.3	0.8-5.21	0 to 4

Table 1.1: Material composition and composition of few DR materials

1.1.3 Permittivity (ϵ_r)

As shown in the table1.1 ϵ_r of DR is in the range between 29 to 80. For low dielectric constant materials performance is likely to be more sensitive to shielding due to increase of field outside the resonator. When ϵ_r is around 40 more than 95% of stored electric energy as well as 60% of the stored magnetic energy are located within the resonator for $TE_{01\delta}$ mode.

1.2 Measurement of coupling and Q-factor

The complete characterization of microstrip line coupled with DR, as shown in figure1.2 and figure1.3, is necessary for the analysis and synthesis of integrated circuits. Ginzton [8] introduced the plotting of loci of various Q-factors. Podcameni [9] suggested a relation for determination of unloaded Q of DR using an indirect method. A simple method is presented in [10] to measure the coupling (β), Q_u, Q_{ex} and Q_L in terms of directly measurable quantities in both reflection and transmission coefficient plane. They also draw the loci for different coupling on smith chart.

However, in this thesis these quantities have been directly measured from the plots of reflection coff. and transmission coff. with respect to frequency. This

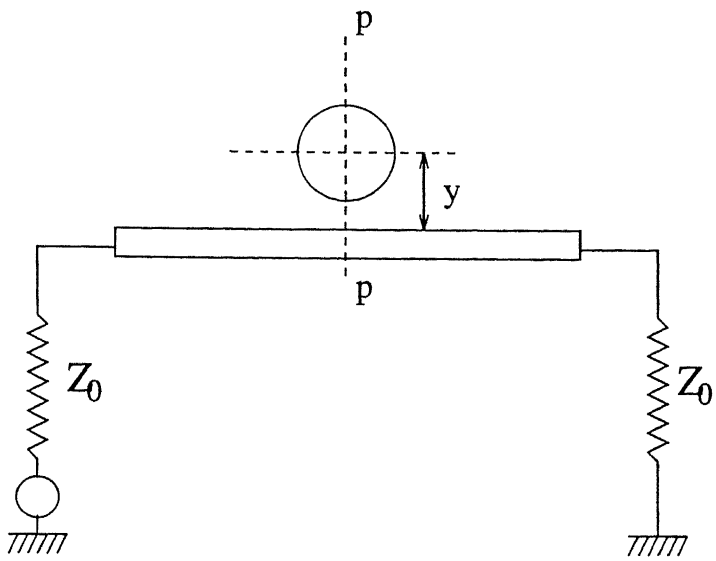


Figure 1.2: DR Coupled to a Microstrip Line

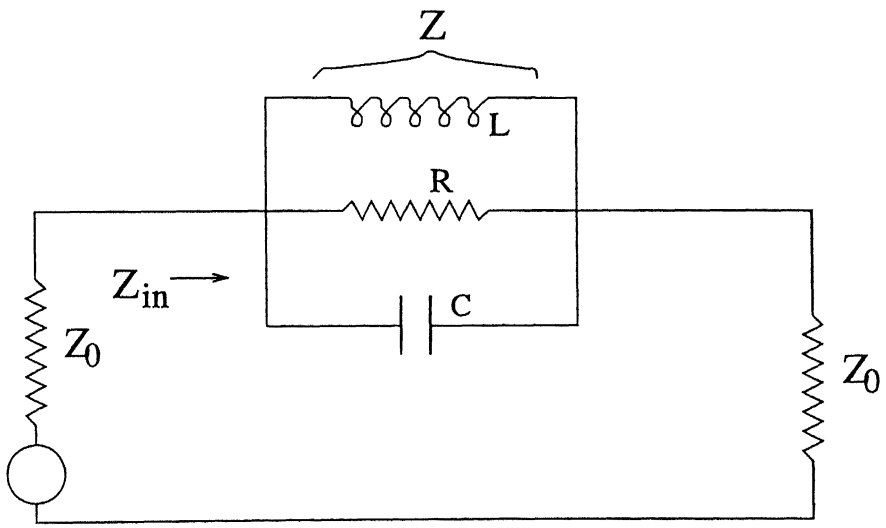


Figure 1.3: Equivalent Circuit of Fig1.2

1.3 Oscillator

Oscillators represent the basic microwave energy source for all microwave systems such as radars, communication, nevigation, and electronic warfare. They can be termed as D.C to R.F converter or an infinite gain amplifier. Typical microwave oscillator essentially have an active device (BJT or FET) and a passive frequency determining element such as a microstrip, SAW, cavity resonator or DR for fixed frequency oscillation and YIG (Yittrioum Iron Garnet) sphere for tunnable oscillator. Microwave oscillator using GUNN or IMPATT diodes dates back to the late 1960s, before that magnetron and klystron were used as sources. Magnetron and klystron are still used for high power application.

Quartz crystal oscillators represent highly stable source, but their operation is limited to few mega Hertz. At microwave frequency these are realised using frequency multiplication. In this process power is increased but at the cost of higher noise figure and less efficiency of device.

With the development of B.J.T and of GaAs MESFET devices at microwave frequency, now transistor based oscillator can be realised at such frequencies. Bipolar oscillator have maximum oscillation frequency which is lower than that of GaAs FET oscillator but later is noiser than the former.

Present trend of oscillator design is mostly emphasised low noise, small size, light weight, high efficiency, temperature stable, low cost and reliable for all fixed frequency oscillation. Tunnable oscillator should additionally have wider band-width, better tunning linearity and reduced settling time.

In this thesis, oscillators are designed using DR as the passive element. Due to its high Q, small size, and excellent integrability in suspended substrate and MIC environment it can be directly used as frequency determining element for realising stable transistor oscillator. Because of its special properties, the transistor dielectric

1.4 Organization of Thesis

The thesis has been divided in five chapters. Chapter 1 is introduction which deals with previous work in this area and procedures used in other chapters of this thesis. Chapter 2 deals with the determination of unloaded quality factor. In chapter 3 experimental procedure have been discussed for measurement of coupling and various Q factors. Chapter 4 discusses the design procedure of oscillator with DR as passive and FET as active element. In the last chapter results and future development have been discussed.

Chapter 2

Determination of Conductor Losses and Unloaded Q-factor

The intrinsic Q-factor of a dielectric resonator, which is determined by the dielectric losses of resonator material, is usually very high. When a DR is shielded by conducting walls in the practical circuits, the unloaded Q-factor of the resonator is degraded due to conductor loss in the top and bottom shielding walls and the dielectric loss in the substrate. Mode matching technique used for analysis of $TE_{01\delta}$ mode for the structure shown in figure2.1, has been given in [11]. The field expression calculated using this technique have been used to compute the conductor losses and conductor Q-factor(Q_c).

2.1 Defination of Q-factor

The Q-factor is assurance of the performance or quality of resonator, is measure of energy loss or dissipation loss per cycle as compared to the stored energy. Q-factor is defined by

$$Q = 2\pi \frac{max.\ stored\ energy}{avg.\ energy\ dissipated\ per\ cycle}$$

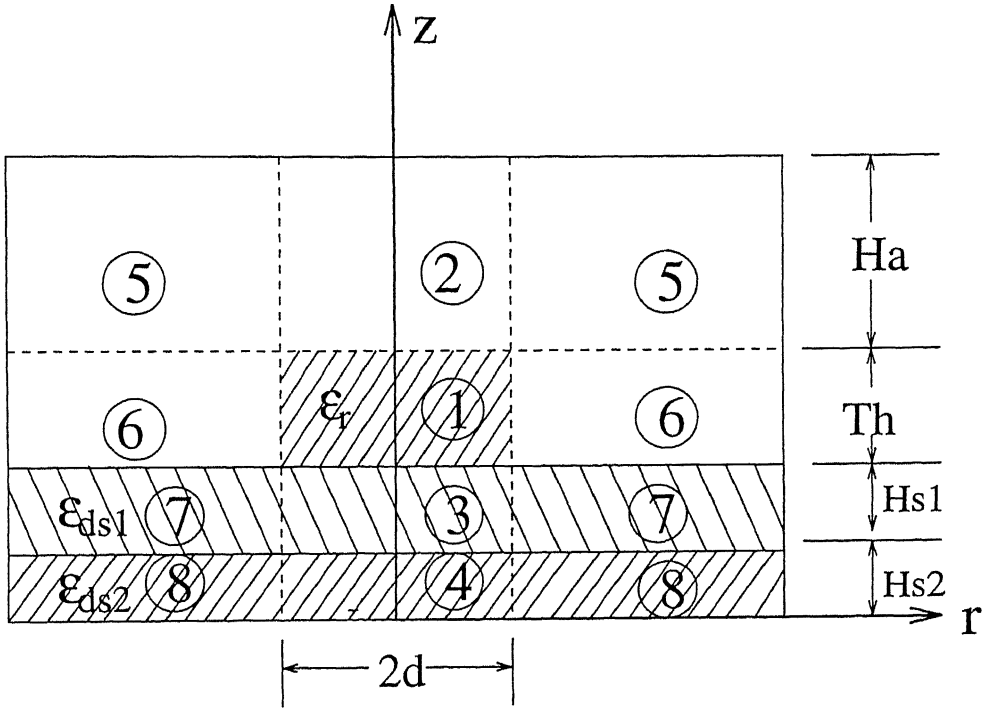


Figure 2.1: Cross sectional view of DR placed in SS environment

Let W_e and W_m are the stored electric and magnetic energy respectively, f_0 is the resonant frequency and P_c is total loss, then Q will be given by

$$Q = 2\pi \frac{(W_e + W_m)}{P_c/f_0} \tag{2.1}$$

EDC (effective dielectric constant) method has been used to evaluate resonant frequency (f_0) of DR in suspended substrte environment [12]. This method requires range in which f_0 lies. Formula to determine lower limit of this range is given in [12], while the upper limit needs to be provided for each set of data. We incorporate a formula for this upper limit.

2.2 Determination of Conductor Loss

then the conductor loss P_c can be expressed as

$$P_c = \frac{1}{2} R_m \int |J_c|^2 \, ds \tag{2.2}$$

Where R_m is real part of surface impedance and given by

$$R_m = \sqrt{\pi f_0 \mu_0 / \sigma} \tag{2.3}$$

Where σ is conductivity of top and bottom ground plate.

2.2.1 Losses in the top plate

In the design of microwave systems containing a DR, generally the top ground plate is much away than the bottom ground plane from DR. So the loss in this plate has lesser effect on overall unloaded Q-factor. Mathematically this loss are represented by

$$P_{c(top)} = \frac{1}{2} R_m \int |J_c|^2 \, ds \tag{2.4}$$

The surface current J_c can be calculated from the boundary condition $J_c = \hat{n} \times \hat{H}$, where $\hat{n}$ is the unit normal vector of the surface. From figure 2.1, it is clear that $\hat{n} = \hat{z}$. Since H_r, H_z and E_ϕ component of fields will be exist for $TE_{01\delta}$ (where δ is slightly less than one) mode in DR, surface current J_c will be contributed only due to H_r. Hence, $J_c = \hat{z} \times H_r$ (region 2 and 5) and from Maxwell's equation H_r is given by

$$H_r = \frac{1}{k_r^2} \frac{\partial^2 H_z}{\partial r \partial z} \tag{2.5}$$

Finally, after satisfying the value of J_c in equation 2.4 we get the expression for $P_{c(top)}$ as follows.

$$P_{c(top)} = \pi R_m \alpha_a^2 B_2^2 . exp(-2\alpha_a d_1)[\frac{I_1}{k_r^2} + \frac{I_2 B_7^2}{k_a^2}] \tag{2.6}$$

Where B_i's, k_r and k_a are the amplitude, radial wave number and axial wave numbers

2.2.2 Losses in the bottom ground plate

Loss in the bottom ground plate play the dominant role in the calculation of unloaded Q-factor. Mathematically this is given by

$$P_{c(bottom)} = \frac{1}{2} R_m \int \left| J_{c(bottom)} \right|^2 ds \tag{2.7}$$

$$J_{c(bottom)} = \hat{z} \times \hat{H}_r(region4, 8)$$

and H_r is given by

$$H_r = \frac{1}{k_a^2} \frac{\partial^2 H_z}{\partial r \partial z} \tag{2.8}$$

The final expression of $P_{c(bottom)}$ is given by

$$P_{c(bottom)} = \pi R_m \alpha_s^2 B_5^2 . exp(-2\alpha_s d_2)[\frac{I_1}{k_r^2} + \frac{I_2 B_7^2}{k_a^2}] \tag{2.9}$$

Where α_a and α_s are the constants and are given in [11].

2.3 Determination of Unloaded Q-Factor

2.3.1 Conductor Quality Factor (Q_c)

With continued progress in the development of DR material, the conductor loss and dielectric loss in the substrate become important factors in determination of overall unloaded Q-factor. Conductor Q-factor is determined using equation 2.1. Expressions for total electric and magnetic energies inside the structure having the DR are given by

$$W_e = \sum_{i=1}^{8} \frac{1}{4} \epsilon_0 \epsilon_r \int |E_{\phi_i}|^2 \, dv = \sum_{i=1}^{8} W_{e_i} \tag{2.10}$$

$$W_m = \sum_{i=1}^{8} \frac{1}{4} \mu_0 \mu_r \int |H_{\phi_i}|^2 , dv = \sum_{i=1}^{8} W_{m_i} \tag{2.11}$$

Where 'i' is the region number.

including DR will be same at a resonant frequency. So complete structure is needed for determination of electric energy required for calculation of Q_c. Which is given by

$$Q_{c(top)} = \frac{2\pi f_0(W_e + W_m)}{P_{c(top)}} \tag{2.12}$$

$$Q_{c(bottom)} = \frac{2\pi f_0(W_e + W_m)}{P_{c(bottom)}} \tag{2.13}$$

Total conductor losses, P_c, are given by

$$\Longrightarrow \quad P_c = P_{c(top)} + P_{(bottom)} \tag{2.14}$$

$$\Longrightarrow \quad \frac{P_c}{2\pi f_0(W_e + W_m)} = \frac{P_{c(top)}}{2\pi f_0(W_e + W_m)} + \frac{P_{c(bottom)}}{2\pi f_0(W_e + W_m)} \tag{2.15}$$

$$\Longrightarrow \quad \frac{1}{Q_c} = \frac{1}{Q_{c(top)}} + \frac{1}{Q_{c(bottom)}} \tag{2.16}$$

$$\tag{2.17}$$

2.3.2 Q-factor Due to Dielectric Loss

Intrinsic Q-factor of the DR (Q_{dr}) is dependent upon on loss tangent of DR materials. Q_d of DR in the suspended substrate environment or MIC environment does not exactly equals $1/tan\delta$. It also depends upon electric energy distribution inside and outside the resonator. A similar argument, is applicable to Q-factor due to dielectric loss in the substrate. By defination Q_d is given by;

$$Q_d = \omega \frac{W}{P_d} \tag{2.18}$$

Where $W = W_e + W_m$ and $P_d = 2\omega W_d tan\delta$

Hence Q-factor due to dielectric loss in different regions are

- Q-factor due to resonator loss is

- Q-factor due to substrate1 (region3 and 7) is

$$Q_{ds1} = \frac{W_e}{(W_{e_3} + W_{e_7})} \frac{1}{tan\delta_{ds1}} \tag{2.20}$$

- Q-factor due to substrate1 (region4 and 8) is

$$Q_{ds2} = \frac{W_e}{(W_{e_4} + W_{e_8})} \frac{1}{tan\delta_{ds2}} \tag{2.21}$$

Here, $tan\delta_{dr}$ and $tan\delta_{ds1,2}$ are the loss tangent of the resonator, substrate1 (region 3 and 7) and substrate2 (region 4 and 8) respectively.

Total dielectric loss(P_{dt}), in the all region is equal to

$$\Longrightarrow \qquad P_{dt} = P_{dr} + P_{ds1} + P_{ds2} \tag{2.22}$$

$$\Longrightarrow \quad P_{dt} = 2\omega W_{e_1} tan\delta_{dr} + 2\omega(W_{e_3} + W_{e_7}) tan\delta_{ds1} + 2\omega(W_{e_4} + W_{e_8}) tan\delta_{ds2} \tag{2.23}$$

$$\Longrightarrow \qquad \frac{P_d}{2\omega W_e} = \frac{W_{e_1} tan\delta_{dr}}{W_e} + \frac{(W_{e_3} + W_{e_7}) tan\delta_{ds1}}{W_e} + \frac{(W_{e_4} + W_{e_8}) tan\delta_{ds2}}{W_e}$$

$$\Longrightarrow \qquad \frac{1}{Q_{dt}} = \frac{1}{Q_{dr}} + \frac{1}{Q_{ds1}} + \frac{1}{Q_{ds2}}$$

$$\Longrightarrow \quad Q_{dt} = \frac{W_e}{W_{e_1} tan\delta_{dr} + (W_{e_3} + W_{e_7}) tan\delta_{ds1} + (W_{e_4} + W_{e_8}) tan\delta_{ds2}} \tag{2.26}$$

For special case of two layer dielectric substrate, Q_{dt} is given by;

1. Suspended substrate case $tan\delta_{ds2} = 0$ (substrate2 is assumed to be air region) and Q_{dt} is

$$Q_{dt} = \frac{W_e}{W_{e_1} tan\delta_{dr} + (W_{e_3} + W_{e_7}) tan\delta_{ds1}} \tag{2.27}$$

2. MIC environment case Q_{dt} can be calculated as follow;

   hs1=hs2=h/2, $\epsilon_{ds1} = \epsilon ds2 = \epsilon ds$

   $tan\delta_{ds1} = tan\delta_{ds2} = tan\delta_{ds}$

$$Q_{dt} = \frac{W_e}{} \tag{2.28}$$

2.3.3 Overall unloaded Q-factor

Once conductor Q-factor and Q-factor due to dielectric losses are calculated, unloaded Q-factor (Q_d) can be calculate as follows.

$$\frac{1}{Q_u} = \frac{1}{Q_c} + \frac{1}{Q_{dt}} \tag{2.29}$$

2.4 Numerical Results

An analytical formulation is developed, to compute the conductor and dielectric loss, and hence Q-factors (Q_c, Q_{dt} andQ_u) in a generalized two layer substrate. Validity of this theory is verified with the data available in literature [13], [1] as the special cases of suspended substrate environment.

The data for Q_d , Q_c and Q_u of DR in suspended substrate environment are generated taking into account of the losses due to substrate, DR and conductor plates.The dielectric loss due to DR is calculated from graph [14] using method of polynomial interpolation.The polynomial is chosen of third degree as $Q_d = a_0 + a_1 f_0 + a_2 f_0^2 + a_3 f_0^3$, where the value of polynomial constant are : a_0=39.75625471 , a_1=-5.112382479, a_2=0.251317681, a_3=-0.003817681.The dielectric loss due to substrate is calculated in same manner from the data given in [15]. Data for Q_u have also been generated for DR in suspended substrate environment, as a function of tuning plate height(Ha) with different parameter such as bottom airgap (Hs2), ϵ_{dr}, ϵ_{ds} etc and they are shown from figure 2.4 to 2.10. These data may be very useful for the design of microwave device containing DR.

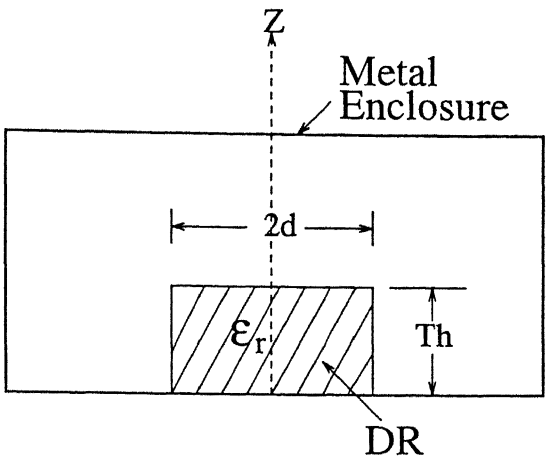


Figure 2.2: DR in Shielded Environment

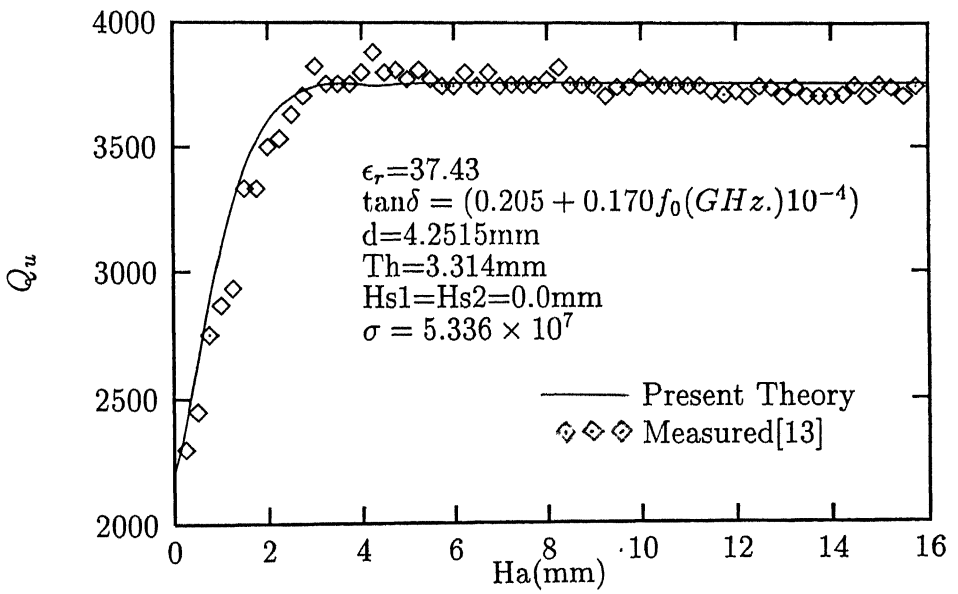


Figure 2.3: Unloaded Q-factor of DR placed in MIC environment

Table2.1

σ=6.14 $\times$ 10^7, Hs1=Hs2=H/2=0.35mm , $\epsilon_{ds1} = \epsilon_{ds2} = 9.6$

ϵ_r	d (mm)	Th (mm)	Ha (mm)	$tan\delta_{ds} \times 10^4$	Q_u [1]	Q_u present Theory
34.19	7.49	7.48	0.72	3.02	2470	2449
34.21	6.995	6.95	1.25	3.19	2440	2435
34.02	5.995	5.98	2.215	3.47	2410	2423
36.13	3.015	4 21	10.10	4.22	1980	2168

Table 2.1: Comparison of present theoretical data for Q_u with that of reported data

2.4.1 Validity of present Theory

A comparison between present theory and experimental data for Q_u available for DR [13] in shielded environment (figure2.2) as special case of DR in suspended substrate environment and is shown in figure 2.3. It shows a good agreement between them for different heights of top plate Ha.

One more comparison is made for DR in MIC environment which is also a special case of suspended substrate environment. Table2.1 shows that comparison. It shows a good agreement with the data reported in [1].

2.4.2 Effect of Bottom Airgap (Hs2)

Figure 2.4 shows the variation of Q_c as function of tunning plate height (Ha) for different Hs2. It is seen from the figure that Q_c increase with the increase of Hs2. This is because of less field interaction with bottom plate which in interms reduces are the conductor loss.

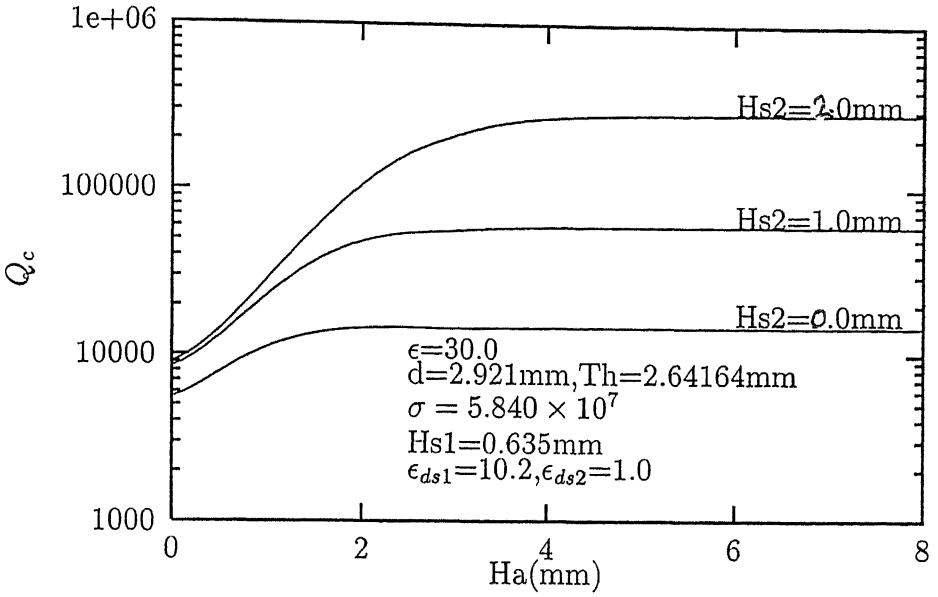


Figure 2.4: Conductor Q-factor of in SS Environment

Q_d decreases due to the fact that the resonant frequency of DR drifts downwards with the increase in Hs2. This causes an increase in electric field distribution in the substrate region which is more lossy than DR, while overall electric energy remain same. Again, the value of conductor Q-factor (Q_c) is much higher than Q_d for fixed value of Ha. This reflect on calculation of Q_u where $1/Q_d$ dominate over $1/Q_c$. Hence similar behaviour is observed for Q_u with increase in Hs2.

2.4.3 Effect of Substrates Parameter on Q_u

Data for Q_u is also generated for various value of permittivity of substrate (ϵ_{ds}) and substrate thickness (Hs1) and is shown in figure 2.6 and 2.7. It is observed from figure 2.6 that the variation of Q_u increases with the increase of Hs1. That is because more field confines to the substrate, so dielectric loss are more. Effect of permittivity of substrate on the Q_u is shown in figure2.7. It is observed from this figure that Q_u is decreases with the increase in ϵ_{ds1}. That is because more and more

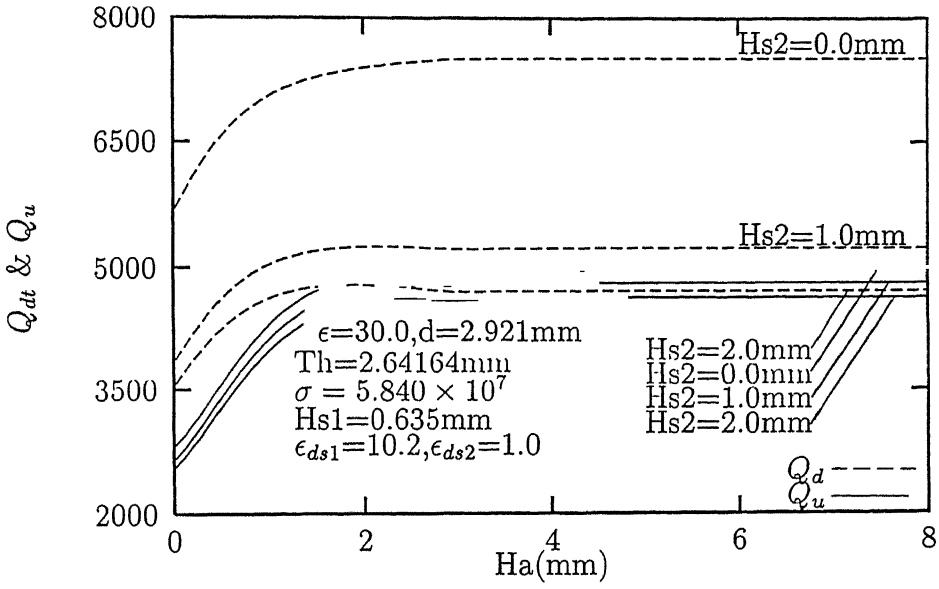


Figure 2.5: Unloaded and dielectric loss Q-factor of in SS Environment

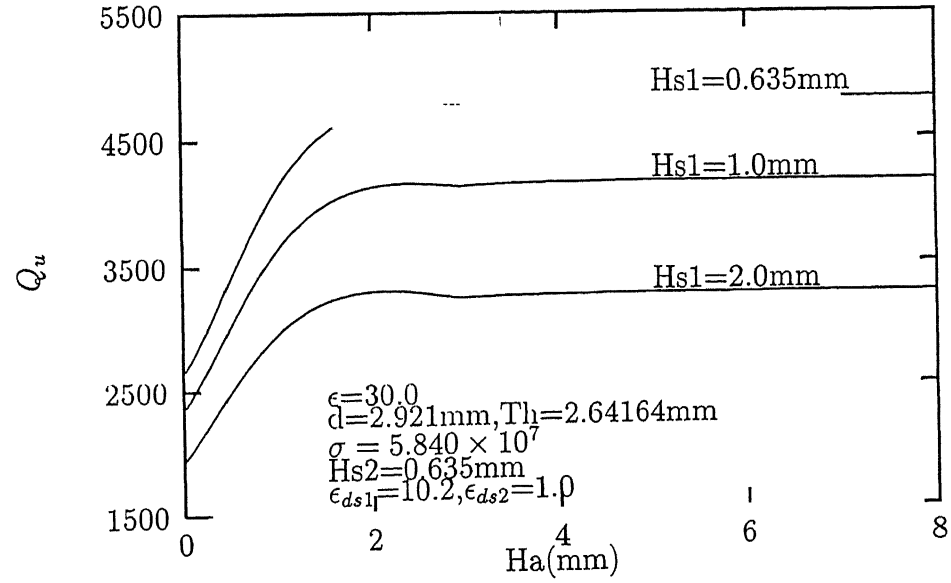


Figure 2.6: Unloaded Q-factor of in SS Environment

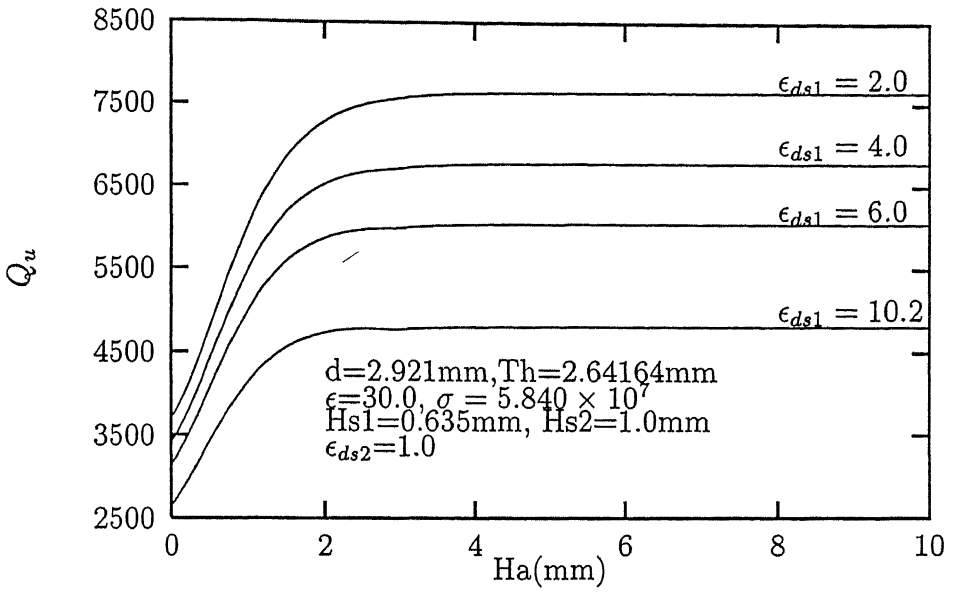


Figure 2.7: Effect of ϵ_{ds1} on Q_u

2.4.4 Effect of DR's Parameter on Q_u

Effect of various parameter of DR on Q_u has also been studies. With increase in DR's radious and height, Q_u increase (figure2.8 and 2.9). That is because the volume of DR is increased in both case and more energy is confine inside the DR rather then in the substrate or in the air region.

As permittivity of DR increases Q_u also increase (figure2.10). As ϵ_r increases, more and more energy will be inside the DR. So their will be less conductor and dielectric loss in the shielding wall and substrate respectively.

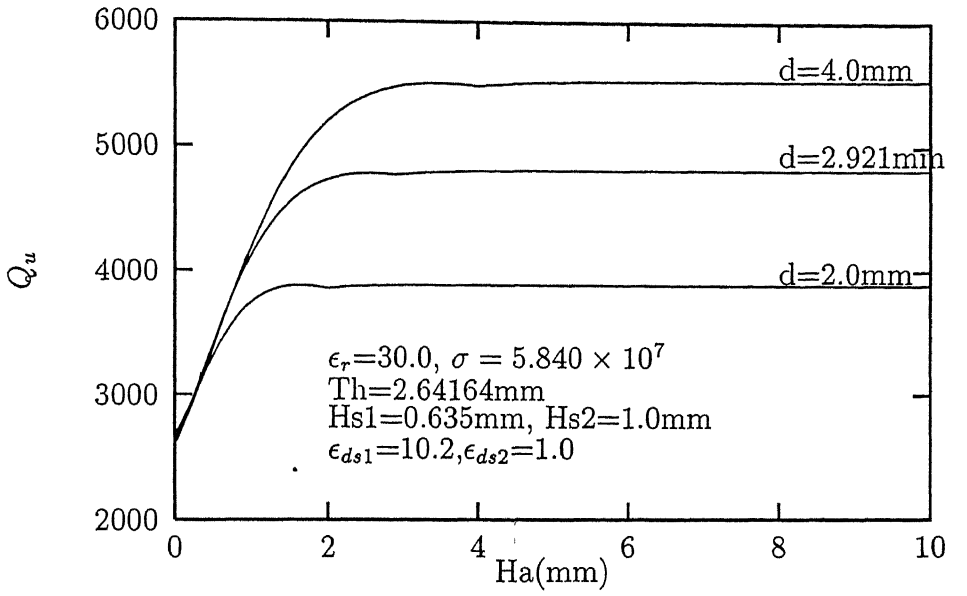
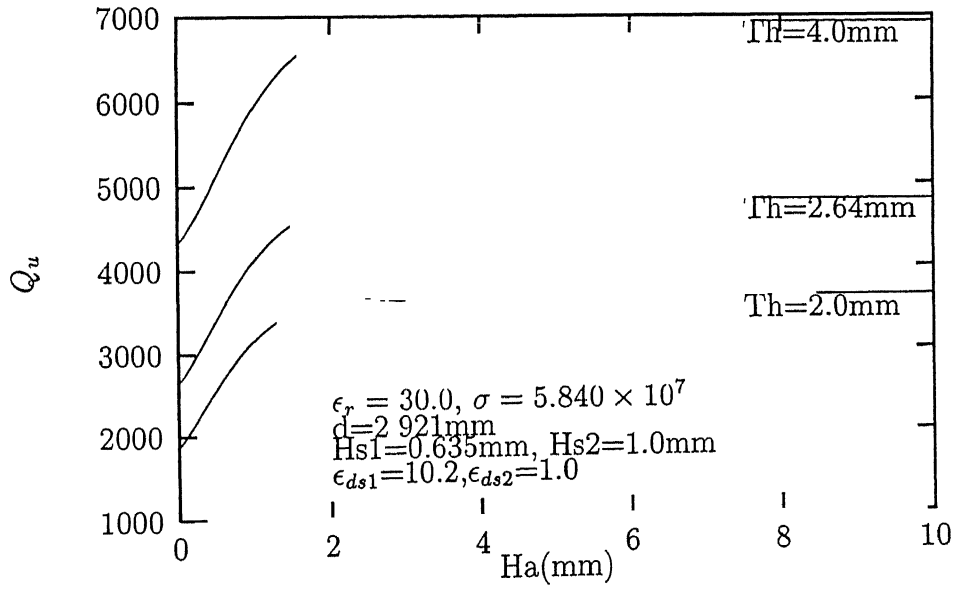


Figure 2.8: Effect of radious of DR on Q_u



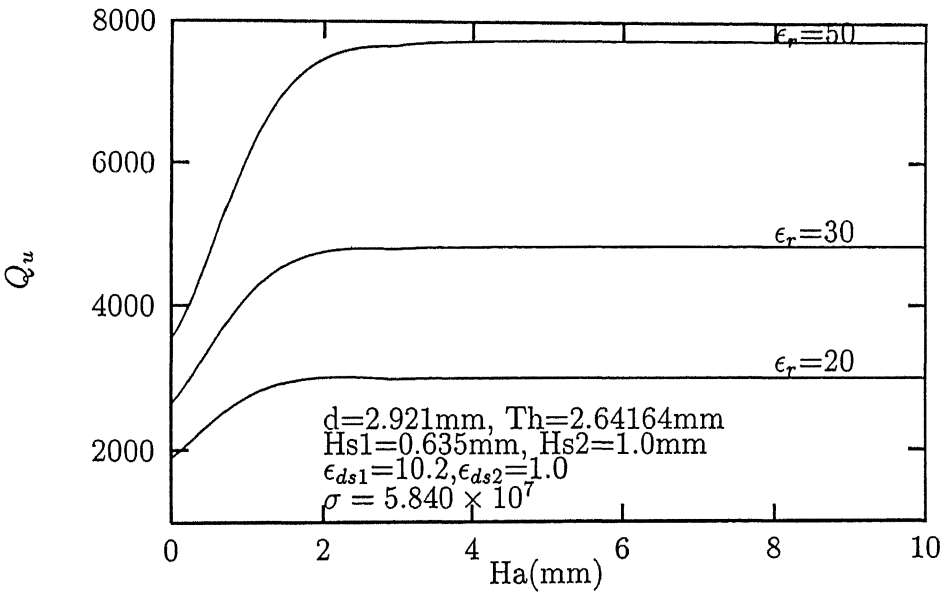


Figure 2.10: Effect of ϵ_r of DR on Q_u

Chapter 3

Measurement Procedure for Determination of Coupling (β), Q_u, Q_{ext} and Q_L

The precise measurement of coupling coefficient (β), resonant frequency (f_0), and Q-factors is necessary and sufficient for complete characterization of DR in suspended substrate or MIC environment for a particular mode. In this chapter, experimental procedure is discussed to measure these quantities for $TE_{01\delta}$ mode.

3.1 Circuit Parameters: An Introduction

At low radio frequencies the simple resonant circuit can be specified completely by stating the circuit parameters in terms of L, C, R_s as shown in figure 3.1. The equivalent description of the microwave resonant circuit cannot be so explicit because as in the waveguide or striplines the ordinary concept of voltage and current does not play its usual role. To define the circuit parameters in any microwave problem

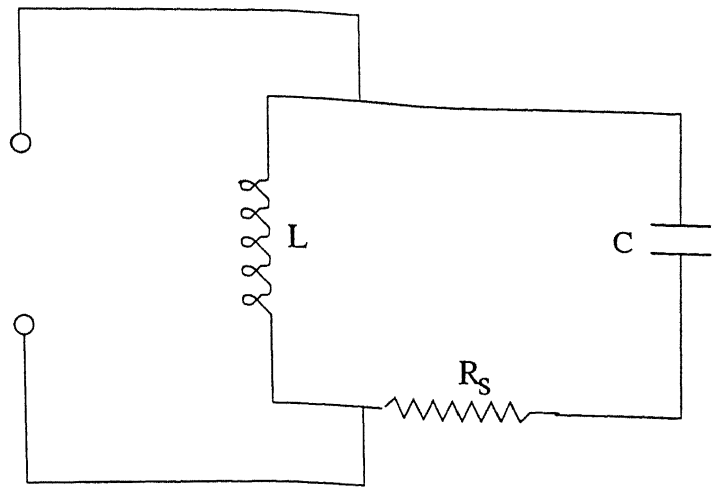


Figure 3.1: Low frequency resonant circuit

in a manner that closely resemble low frequency circuit practice. However, the microwave equivalent circuit concept is complicated by the relatively close spacing of the resonant frequencies of the microwave cavity. In most practical situation the cavity or resonator are used at sufficiently low frequencies so that only one mode is excited at a time, thus making it possible to represent the energy stored in the field of some particular mode by the energy stored in the lumped parameters of the equivalent circuit.

A consideration of low frequency analogue clarifies the meaning of equivalent circuit parameters. The three parameters shown in figure 3.1 can be related to three universally useful relations.

$$\omega_0^2 = \frac{1}{LC} \tag{3.1}$$

$$Q_0 = \frac{\omega_0 L}{R_S} \tag{3.2}$$

$$R_0 = \omega_0 L Q_0 = \frac{(\omega_0 L)^2}{R_S} \tag{3.3}$$

Where R_0 is shunt resistance.

The three quantities defined by these relations can be measured experimentally

resulting in

$$L = \frac{R_0}{\omega_0 Q_0} \tag{3.4}$$

$$C = \frac{Q_0}{\omega_0 R_0} \tag{3.5}$$

$$R_S = \frac{R_0}{Q_0^2} \tag{3.6}$$

The microwave resonant cavity differs that from low frequency circuit in two respects: firstly, the equivalent circuit parameters must be established separately for each mode under consideration and secondly, the quantity R_0 is not uniquely defined due to ambiguity in the meaning of voltage and currents.

3.2 Coupling Factor of DR Coupled to Microstrip

In order to effectively use DR in the microwave circuits, it is necessary to have an accurate knowledge of the coupling between DR and transmission line. The $TE_{01\delta}$ mode of the cylindrical resonator can be easily coupled with microstrip, fin line, magnetic loop and waveguides.

Figure 3.2 shows the magnetic coupling between DR and microstrip line. The resonator is placed on the microstrip. The lateral distance 'y' between DR and microstrip primarily determines the amount of coupling between line and DR. $TE_{01\delta}$ mode in DR can be approximated by a magnetic dipole of moment M. The coupling between the line and the resonator accomplished by orienting the magnetic moment of the resonator, perpendicular to microstrip plane so that the magnetic lines of the resonator link with those of microstrip line, as shown in figure 3.2. The dielectric resonators placed adjacent to the microstrip line operates like a reaction cavity that reflects the RF energy at the resonant frequency. The equivalent circuit of the resonator coupled to microstrip line is shown in figure 3.3.

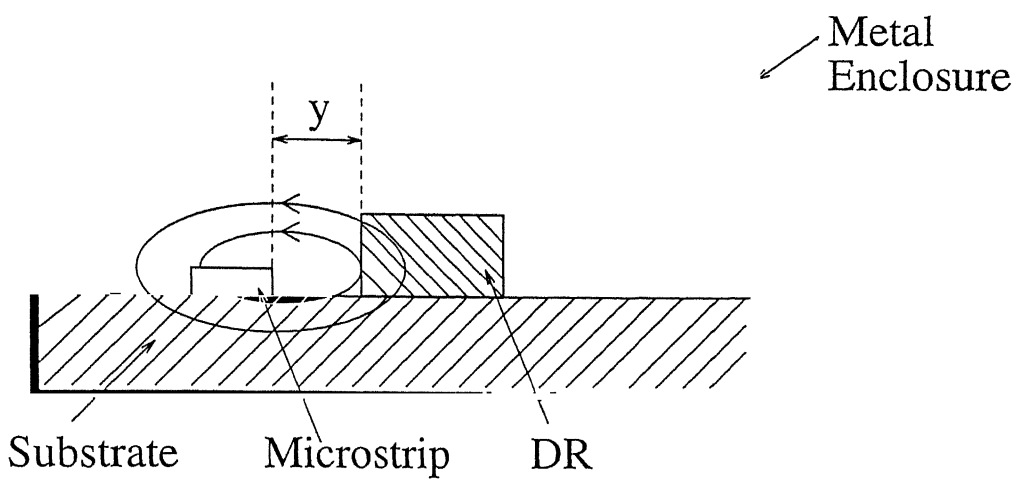


Figure 3.2: Coupling between DR and Microstrip

In figure 3.3 L_r, C_r and R_r are the equivalent parameters of the dielectric resonator, L_1, C_1 and R_1 are the equivalent parameters of the microstrip line, and L_m characterizes the magnetic coupling. The resonator impedance Z in series with the transmission line is easily determined to be

$$Z_{in} = j\omega L_1 + \frac{\omega^2 L_m^2}{R_r + j\omega(L_r - 1/\omega^2 C_r)} \tag{3.7}$$

Around resonant frequency ωL_r is negligible, so Z_{in} is given by

$$Z_{in} = \omega . Q_u \frac{L_m^2}{L_r} . \frac{1}{1 + jX} \tag{3.8}$$

Where $X = 2Q_u(\Delta\omega/\omega)$ and unloaded Q-factor, is given by

$$Q_u = \frac{\omega_0 L_r}{R_r} \tag{3.9}$$

$$\omega_0 = \frac{1}{\sqrt{L_r C_r}} \tag{3.10}$$

And S-matrix correspond to figure 1.3 (chapter1) will be

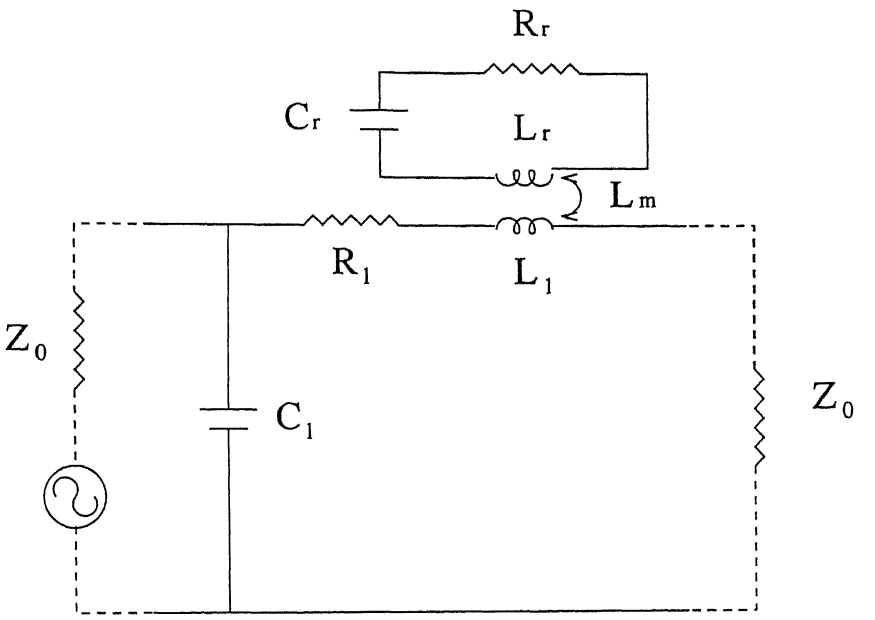


Figure 3.3: Equivalent circuit of DR coupled to microstrip

At the resonance

$\Delta\omega = 0$, so X=0 and

$$Z_{in} = R = \omega_0 Q_u \frac{L_m^2}{L_r} \tag{3.11}$$

$$S = \begin{bmatrix} \frac{\beta}{\beta+1} & \frac{1}{1+\beta} \\ \frac{1}{1+\beta} & \frac{\beta}{\beta+1} \end{bmatrix}$$

The equivalent circuit of DR coupled with microstrip or suspended strip line can also be represented as resonant circuit shown in figure 1.3. L, C, R satisfy the following equations.

$$L = \frac{L_m^2}{L_r} \tag{3.12}$$

$$C = \frac{L_r}{\omega_0^2 L_m^2} \tag{3.13}$$

$$R = \omega_0 Q_u \frac{L_m^2}{L_r} \tag{3.14}$$

From equation 3.12 to 3.15, quantities L, R, and C can be represented in terms of directly measurable parameter β, ω_0, and Q_u as given below.

$$R = \omega_0 Q_u \frac{2\beta Z_0}{\omega_0 Q_u} = 2\beta Z_0 \tag{3.16}$$

$$L = \frac{R}{\omega_0 Q_u} \tag{3.17}$$

$$C = \frac{Q_u}{\omega_0 R} \tag{3.18}$$

3.3 Measurement Setup

Measurement is performed on the Trans-Tech resonator made of D8600 material (ϵ_r=37.43). Same resonator used in the design of oscillator. The resonator is placed on the dielectric substrate and on the plane of symmetry pp' as shown in figure 3.4. Now this is a two port symmetric and reciprocal device. The substrate is taconic plastic (RT duroid) of thickness of 0.8mm and of relative dielectric constant is 2.18. The characteristic impedance of microstrip line is 50Ω (conductor width 2.4mm). The spacing between the resonator and microstrip line (y) is gradually increased in the step of 1mm.

Great care is exercised in measurement of coupling. The first and very important step of the measurement procedure is to measure the scattering parameter (Reflection and Transmission Coefficient) of this two port symmetric device using a Hewlett Packard HP8410 network analyser. About 20-25 point around the resonance are measured for each spacing between DR and microstrip line.

The reflection coefficient($S_{11} = S_{22}$) and transmission coefficient($S_{21} = S_{12}$) measured for three different spacing between DR and microstrip line, are plotted as shown in figure 3.5 and figure 3.6. Both plot shows three peaks/troughs, among these peaks/troughs first will be due to fundamental mode $TE_{01\delta}$ because of the

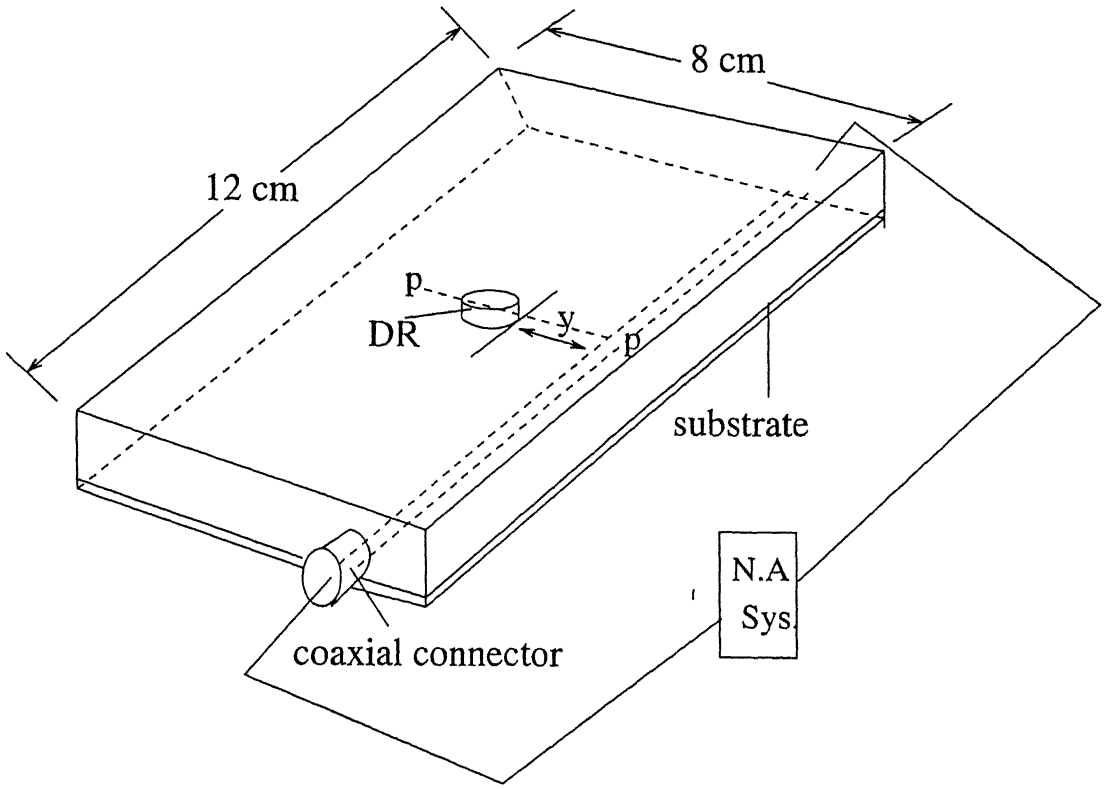
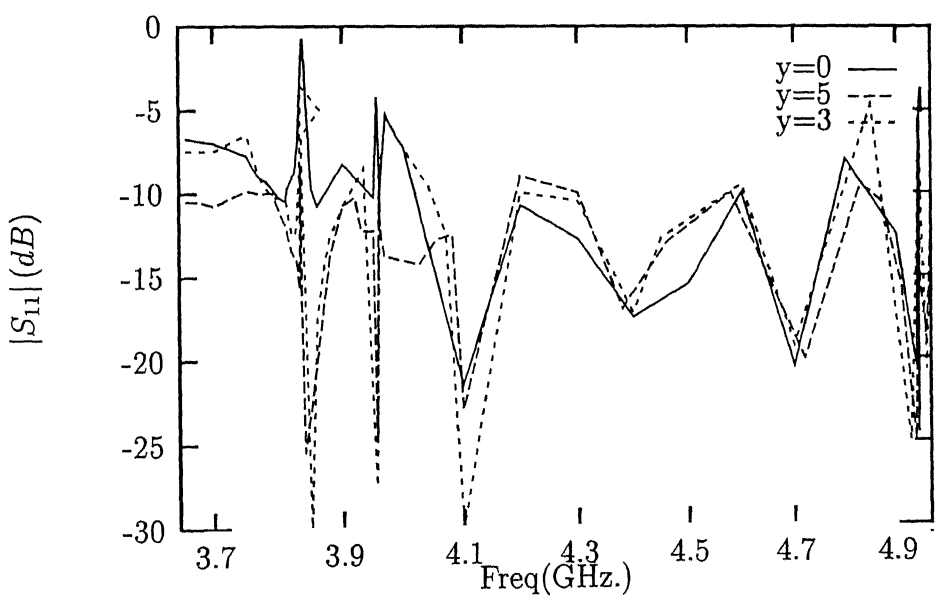


Figure 3.4: Experimental setup



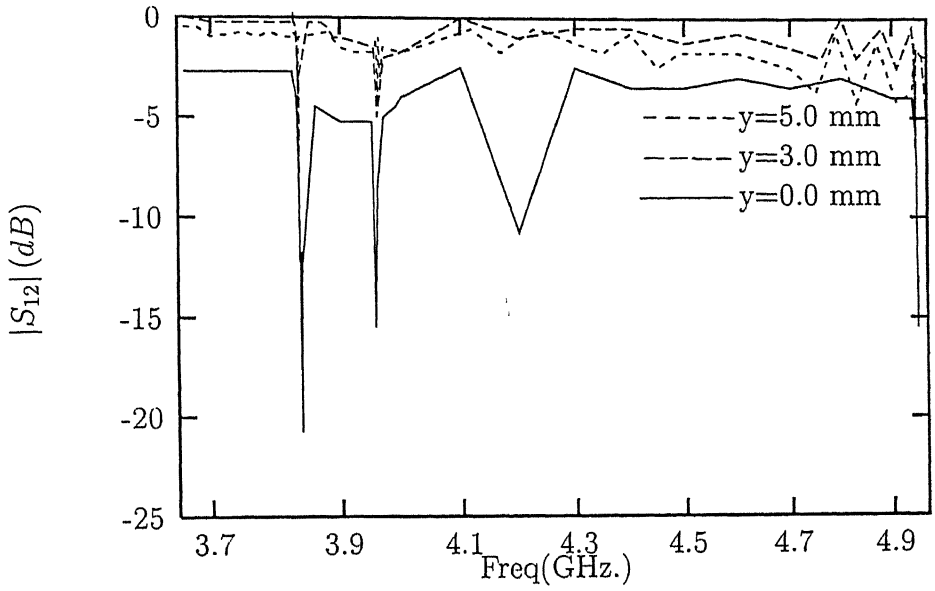


Figure 3.6: Measured S_{12} for different spacing 'y'

respond to fundamental mode.

- It has been observed that in the absence of top conductor plate second peak vanished. It also observed that second peak is generally not symmetric about resonance. Hence second peak is due to reflection.

- Third peak occurs around 5 GHz with less amplitude compare to first peak so it must be because of higher order mode.

Figure 3.7 and 3.8 shows the measured reflection and transmission coefficient around resonance. It is observed that resonant frequency drift slightly with the spacing y.

3.4 Calculation for β, Q_L, Q_u and Q_{ext}

The coupling coefficient (β) can be calculated by either reflection coefficient or trans-

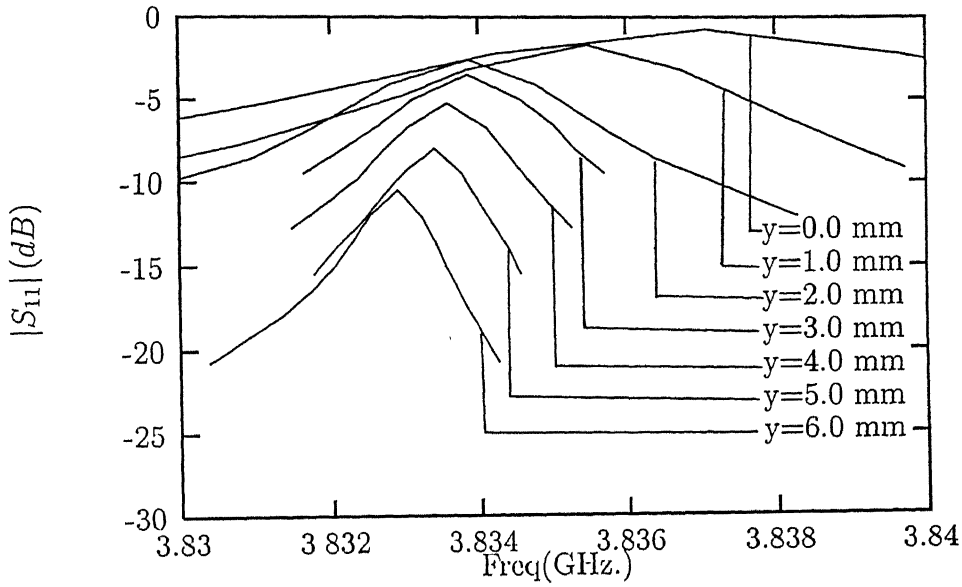
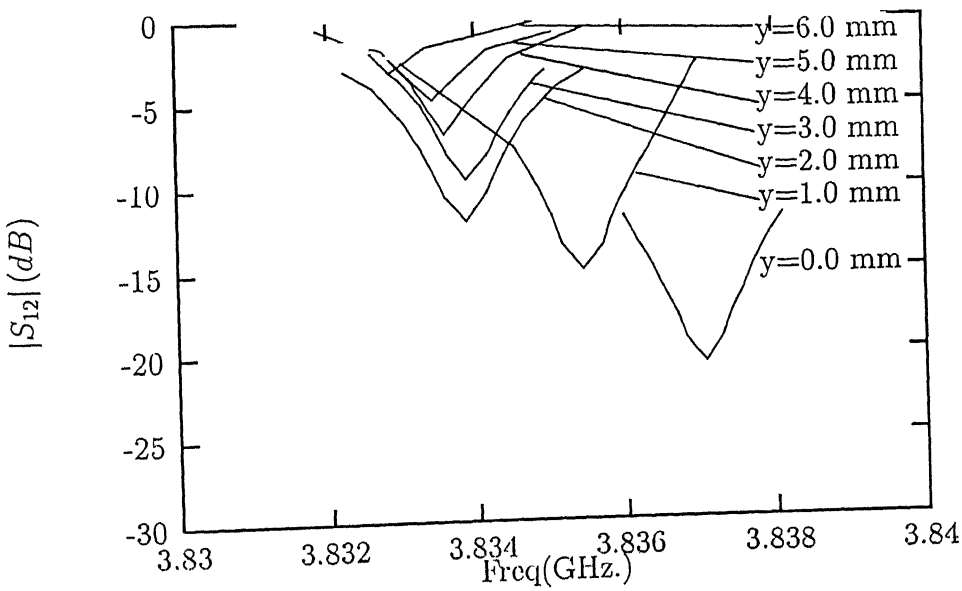


Figure 3.7: Measured S_{11} around resonance for different spacing 'y'



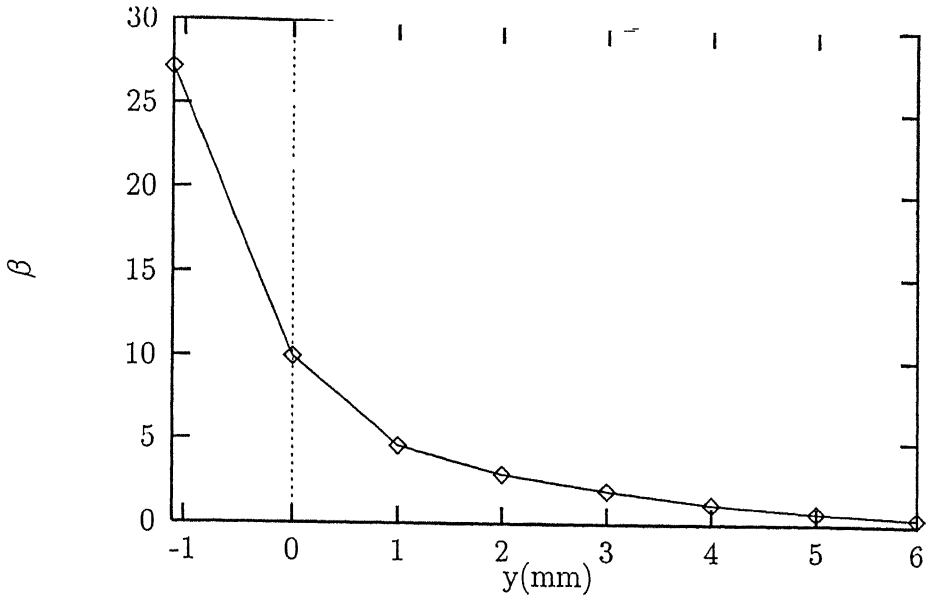


Figure 3.9: Coupling between DR and microstrip

Where S_{110} and S_{210} are real quantities representing the reflection and transmission coefficients respectively at the resonant frequency.

Figure 3.9 shows the variation of coupling with the spacing y. The critical coupling($\beta = 1$) occurs when the power dissipation in the resonator is equal to the power dissipated in external circuit. The critical coupling corresponds to $S_{110} = S_{210}$=0.5.

As shown in figure 3.10, Q_L is 3dB point on reflection coefficient plane can be measured directly. from the reflection coefficient plane. Q_L is 3dB point on reflection coefficient plane.

$$Q_L = \frac{f_0}{f_1 - f_2}$$

Q_L can also be calculated from transmission coefficient plane as follows [16];

$$L_0(dB) = 20 \log S_{210} \tag{3.20}$$

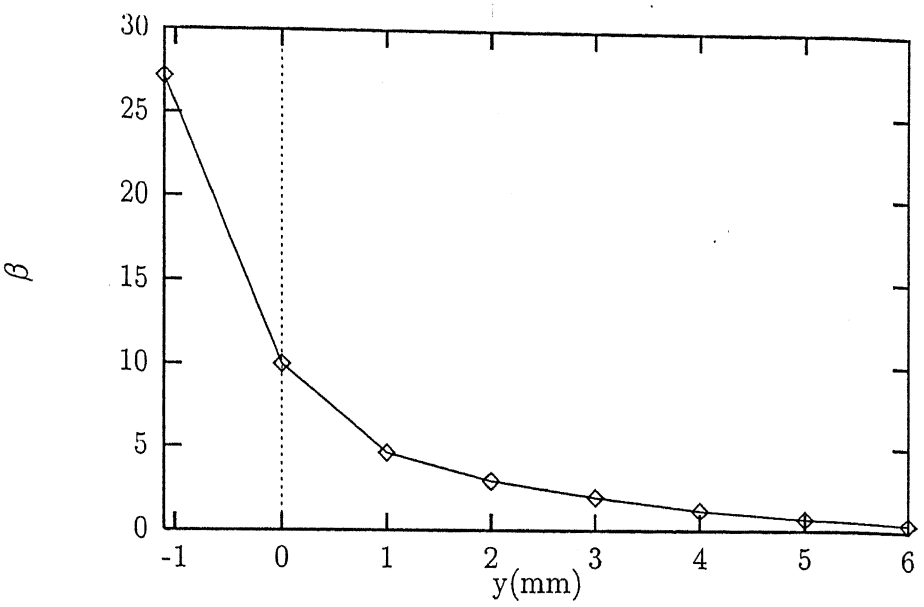


Figure 3.9: Coupling between DR and microstrip

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$$Q_L = \frac{f_0}{f_1 - f_2}$$

Q_L can also be calculated from transmission coefficient plane as follows [16];

$$L_0(dB) = 20 \log S_{210} \tag{3.20}$$

$$X(dB) = 3 - 10 \log[1 + 10^{-0.1L_0}] \tag{3.21}$$

Once X is calculated Q_L can be calculated from transmission coefficient plane (transmission coefficient Vs. frequency plot), as shown in figure 3.11. Since Q_L is depen

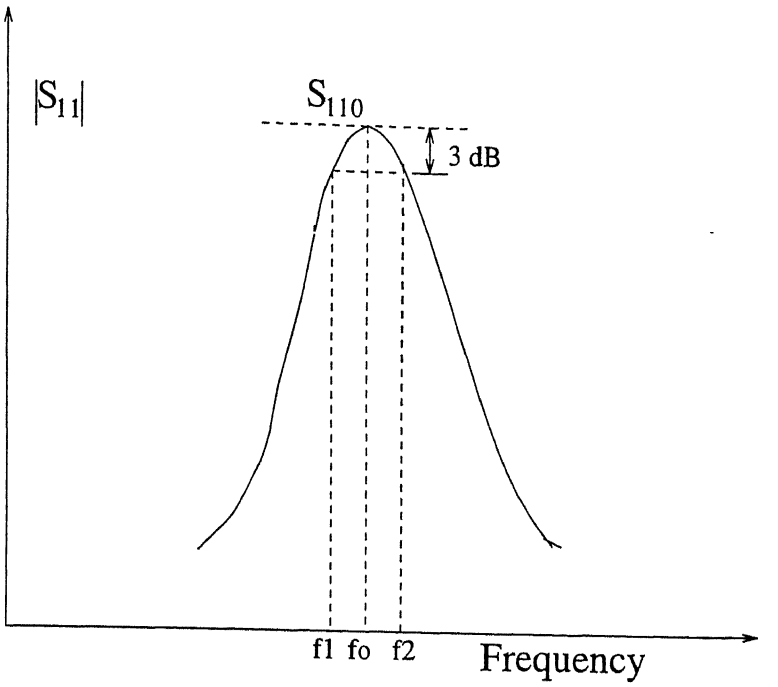

Figure 3.10: Definition of various terms in reflection plane

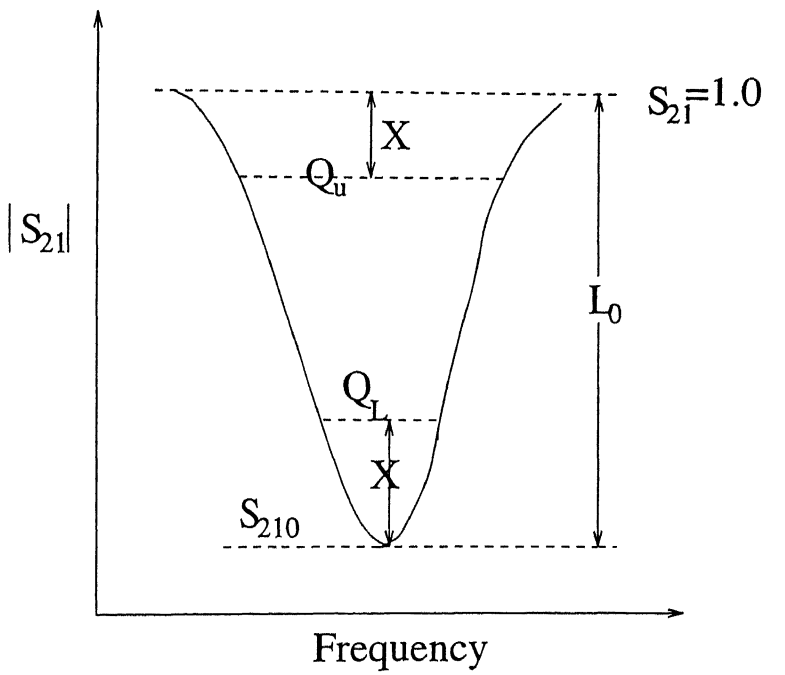

Figure 3.11: Definition of various terms in transmission plane

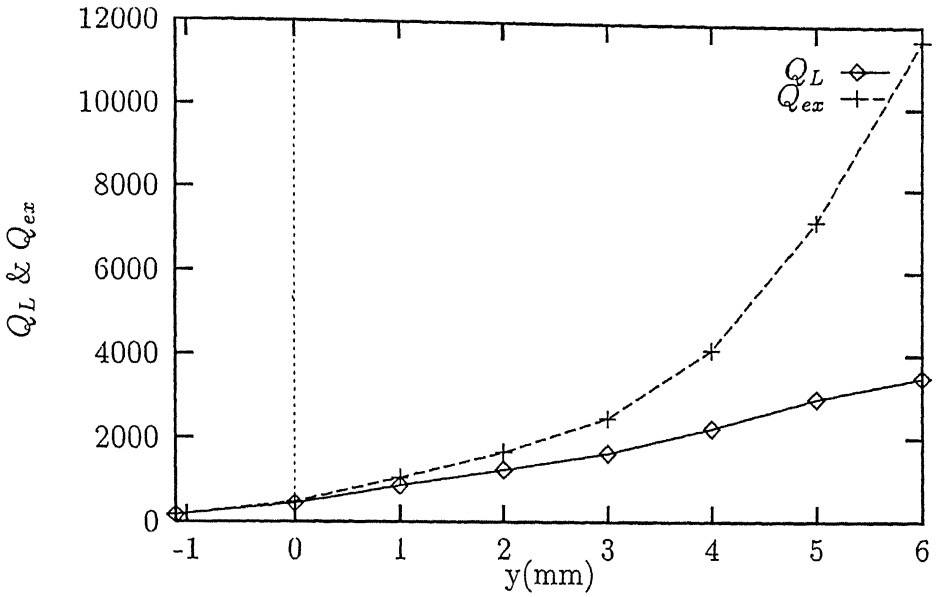

Figure 3.12: Variation of Q_L and Q_{ext} of Loaded DR

dent on spacing y, so generally X is an odd figure; e.g 2.87,2.674 etc, and, hence, it is difficult to find this point (X) accurately on transmission coefficient plane. So it is better to calculate Q_L in reflection coefficient plane.

Once β, f_0, and Q_L are calculated the other Q-factors, Q_u and Q_{ext} can be calculated as follow;

$$Q_{ext}\beta = Q_u = Q_L(1+\beta)$$

The variation of Q_{ext} and Q_L are plotted with spacing y and is shown in figure 3.12. Table3.1 shows Q_u and coupling (β) for different spacing between resonator and microstrip. The difference between the largest and smallest value of Q_u is less than 3%. That is, the Q_u of the covered DR is supposed to be depend only on the distance from resonator to the top and bottom conductors, and not on the lateral position of DR. So it should be almost constant.

y(mm)	β	Q_u	y(mm)	β	Q_u
-1.1	27.12	4848	3.0	1.99	4919
0.0	9.99	4880	4.0	1.19	4906
1.0	4.62	4880	5.0	0.69	4977
2.0	2.95	4818	6.0	0.43	4908

Table 3.1: Data for coupling and unloaded Q-factor of DR coupled with microstrip

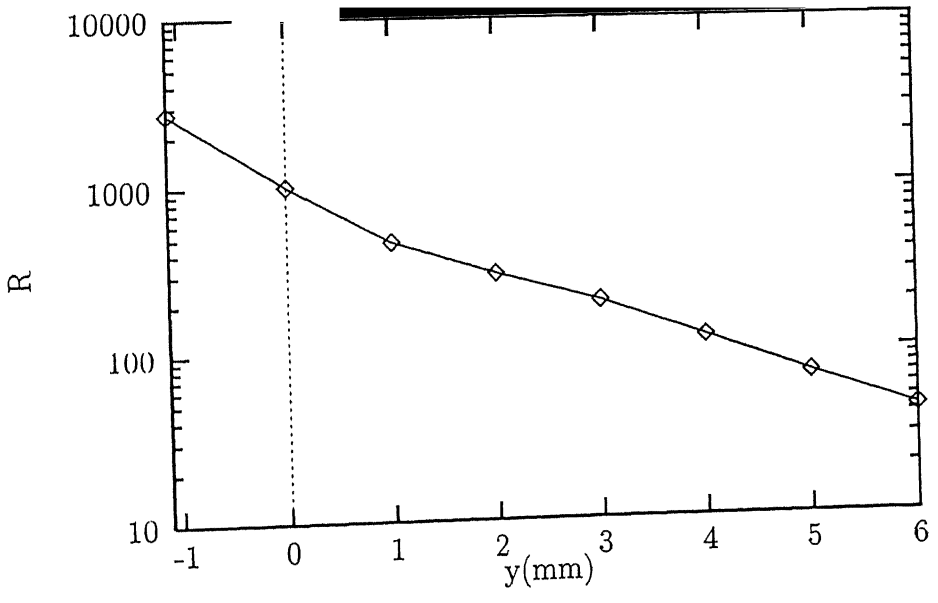

Figure 3.13: Equivalent resistance of DR coupled to microstrip

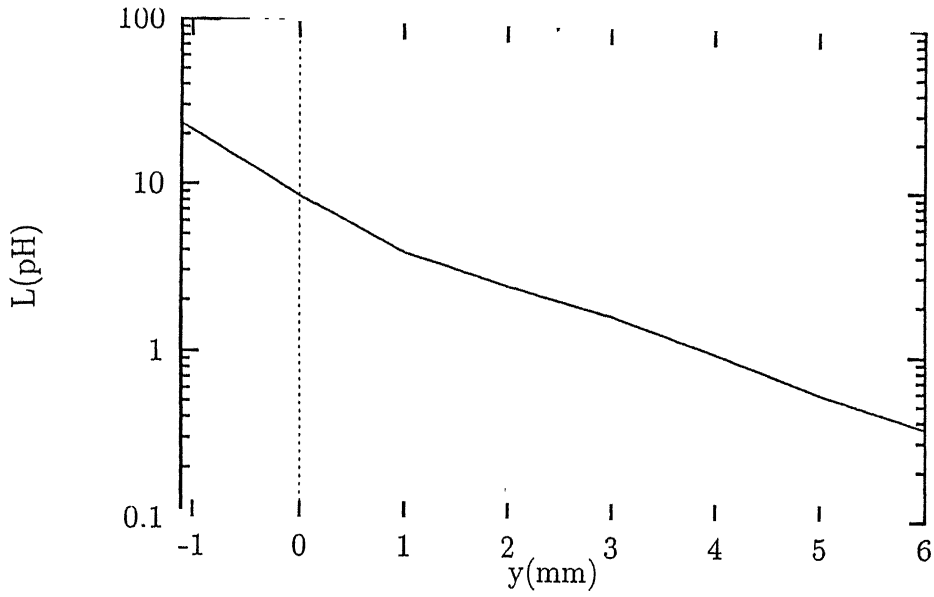

Figure 3.14: Equivalent inductance of DR coupled to microstrip

3.5 Results

The equivalent circuit parameter (R, L, and C) of the DR coupled to microstrip, on plane pp' as shown in figure 1.2, are calculated directly from measurable quantity β, f_0 and Q_u. These parameters are plotted against lateral spacing of DR from microstrip as shown in figure 3.13 to figure 3.15. All these plot might be very useful in the design of microwave system containing the DR, since these give the direct value of the circuit parameters. It is observed from the figure 3.13 that R is decreasing with the increase of spacing y. As spacing y increases lesser power will be coupled to DR so that dielectric loss is lesser. Hence R is going downwards with the increase in spacing y. Again, it is seen from figure 3.11 that the equivalent inductance L($=L_m^2/L_r$) also decrease with increase in y. This is because of lesser mutual coupling (L_m) taking place when y increases. The equivalent capacitance increase with the increase in spacing y is shown in figure 3.15. Here, increase in equivalent capacitance is due to fact that mutual coupling between DR and microstrip (L_m) decreases with the increase in spacing (equation 3.13).

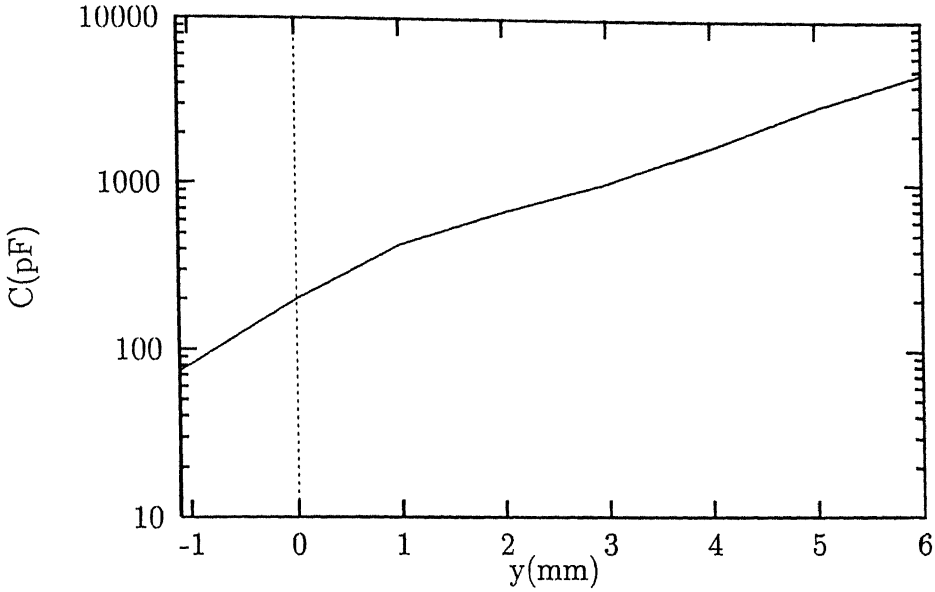

Figure 3.15: Equivalent capacitance of DR coupled to microstrip

Chapter 4

Design and Fabrication of DR Oscillator

The transistorized oscillators can be classified according to whether their design is based on linear device(BJT or FET) parameters such as the S- and Y-parameters, or based on large signal behavior of the transistor. The former category is some times known as the linear oscillator where as the later is sometimes referred to as the power or large signal oscillator.

4.1 Theory of Oscillation

A simple and popular method of oscillator to place a tunning resonator across the input terminals of two port network is shown in figure 4.1. The condition for oscillation for such oscillator can be expressed as [17]

$$K < 1 \tag{4.1}$$

$$\Gamma_1 S'_{11} = 1 \tag{4.2}$$

If and only if

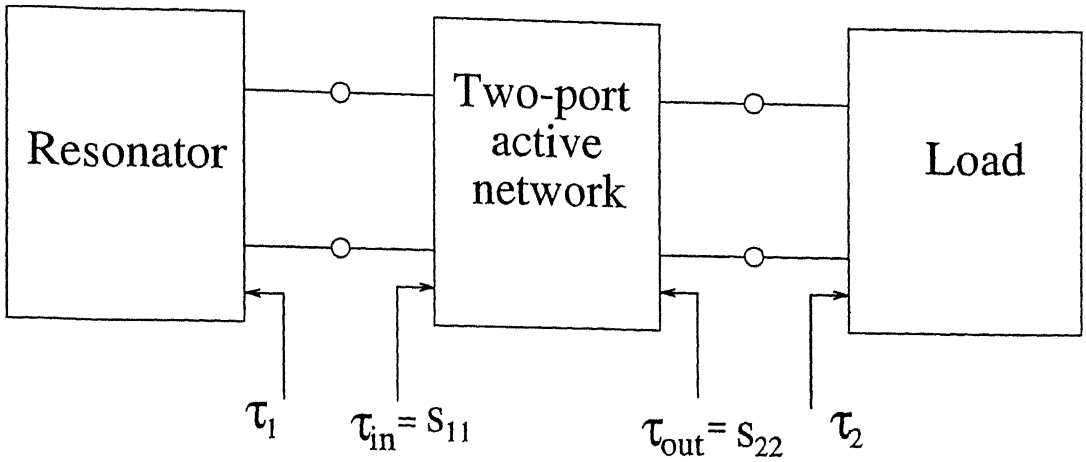

Figure 4.1: Block diagram of an oscillator [17]

In order to start and sustain the oscillations it is necessary that

$$|\Gamma_1| > \left|\frac{1}{S'_{11}}\right| \tag{4.4}$$

S'_{11} and S'_{22} are given by

$$S'_{11} = S_{11} + \frac{S_{12}S_{21}\Gamma_2}{1 - S_{22}\Gamma_2} \tag{4.5}$$

$$S'_{22} = S_{22} + \frac{S_{12}S_{21}\Gamma_1}{1 - S_{11}\Gamma_1} \tag{4.6}$$

The stability factor K, known as Rollet stability factor is defined as

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2\,|S_{21}S_{12}|} \tag{4.7}$$

Where $\Delta = S_{11}S_{22} - S_{21}S_{12}$, and S_{ij} are S-parameters of the device.

The equation 4.2 to 4.4 are equivalent to Barkhausen criterion for oscillator design.

Equation 4.2 can also be written as $R_{in} + R_1 = 0$ *and* $jX_{in} + jX_1 = 0$ Or

$$R_{in} = -R_1 \quad and \quad X_{in} = -X_1 \tag{4.8}$$

Where R_{in}, R_1, and X_{in}, X_1 are shown in figure 4.2.

The stability factor K of the active network should be less than unity in order to

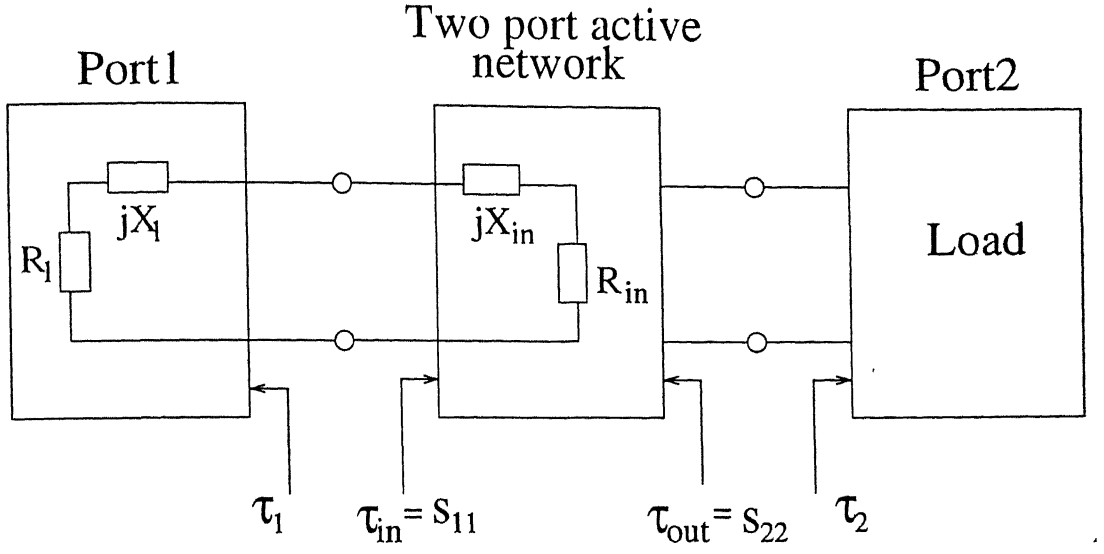

Figure 4.2: Oscillation in port 1 [17]

a bypass capacitor) should be added. The positive termination Γ_1 and Γ_2 must be designed in such a way that the input and output ports are resonating simultaneously at the same frequency.

4.2 Oscillator Design

In the design of stable oscillator, DRs are generally used as frequency determining circuit, feedback circuit, or a matching circuit element. This reduces the size as well as cost provided we should be able to eliminate undesirable features of mode jumping and hysteresis. A stable TDRO has a high efficiency circuit of more simple construction, making it an obvious choice.

An oscillator circuit can be represented as either a series or parallel resonant circuit. The commonly used table TDRO's can be divided in two part: one using the DR as series element, and other using the DR as parallel feedback element. In this thesis, the oscillator is designed using former type. Figure 4.3 shows the basic circuit diagram of a series feedback transistor DRO.

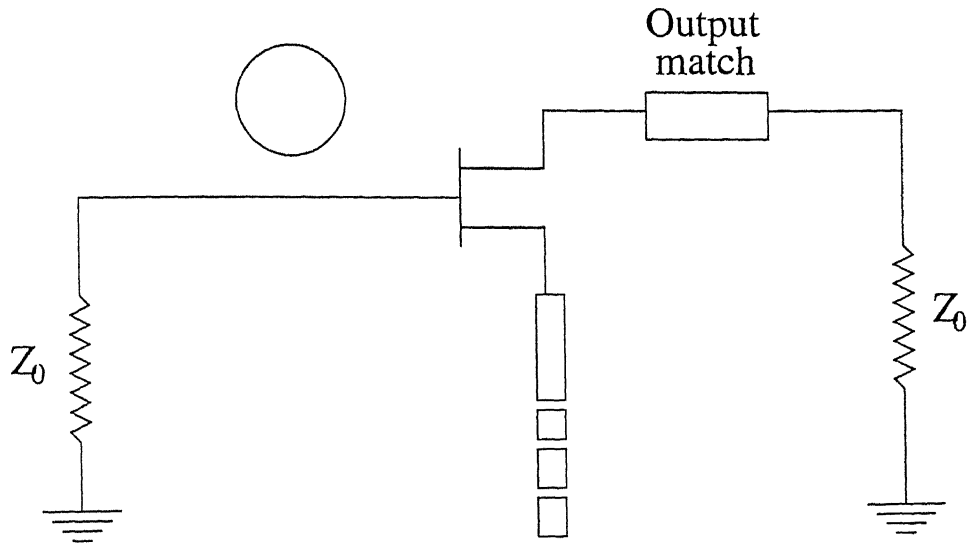

Figure 4.3: DR as a series feedback element in DRO [16]

4.3 Choice of transistor and its biasing

4.3.1 Choice of transistor

In the design of any active circuit, choice of transistor (or any other active element) is very critical. Because it decide the maximum available gain and noise figure of the device.

So we first select a transistor of desired capability and frequency range. A FET NE70083 [NEC] is selected for oscillator design with following specification at 4GHz.

Noise Figure (NF)=0.7 dB

Max. available Gain(G)=14.0dB

Output power (Po)=14.5dBm

Power dissipated (Pd)=400 mW

And S-parameter at $V_{ds} = 3V$ and $I_{ds} = 30mA$ are

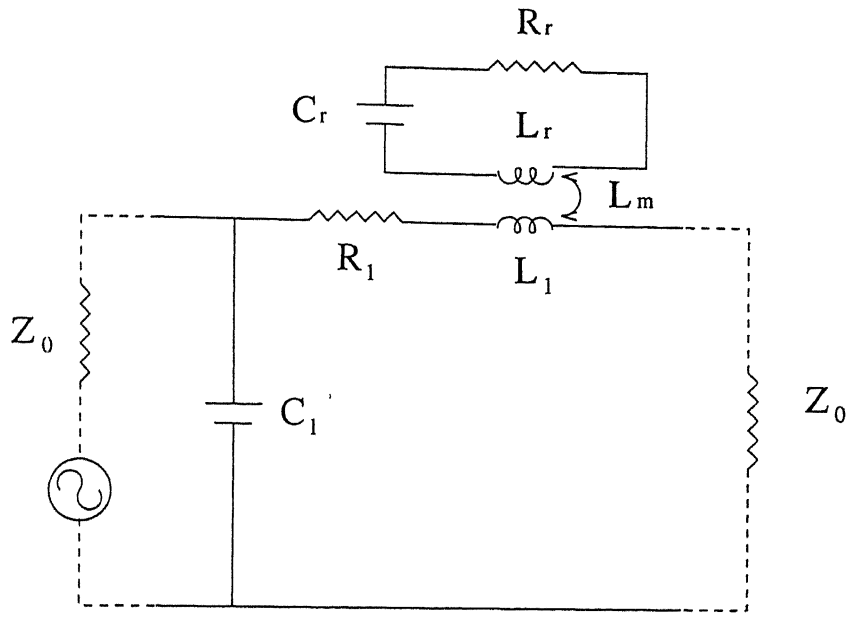

Figure 4.4: Device characteristic

4.3.2 Biasing of the transistor

For the design of biasing circuit we fixed the Q-point, from device characteristic, which decide available gain and actual S-parameter. Figure 4.4 shows the Q-point selected as $V_{ds} = 3V$ and $I_{ds} = 30mA$. At this point S-parameters are given above. Figure 4.5 shows the circuit diagram of oscillator. The value R, L, C_1 and C_2 are calculated as follow.

C_1 and C_2 must act as short-circuit at 4.0 GHz. and as a open circuit at dc.

$$\frac{1}{\omega C} \cong 10(\text{say}) \text{ at 4 GHz.}$$

$$\frac{1}{\omega C} \approx \infty \text{ at dc}$$

$$C_1 \text{ and } C_2 \approx 0.1 \text{ nF} = 100\text{pf}$$

Inductor (L) should be act as a short circuit for dc and o/c at 4 GHz., so that no output signal should leak to ground.

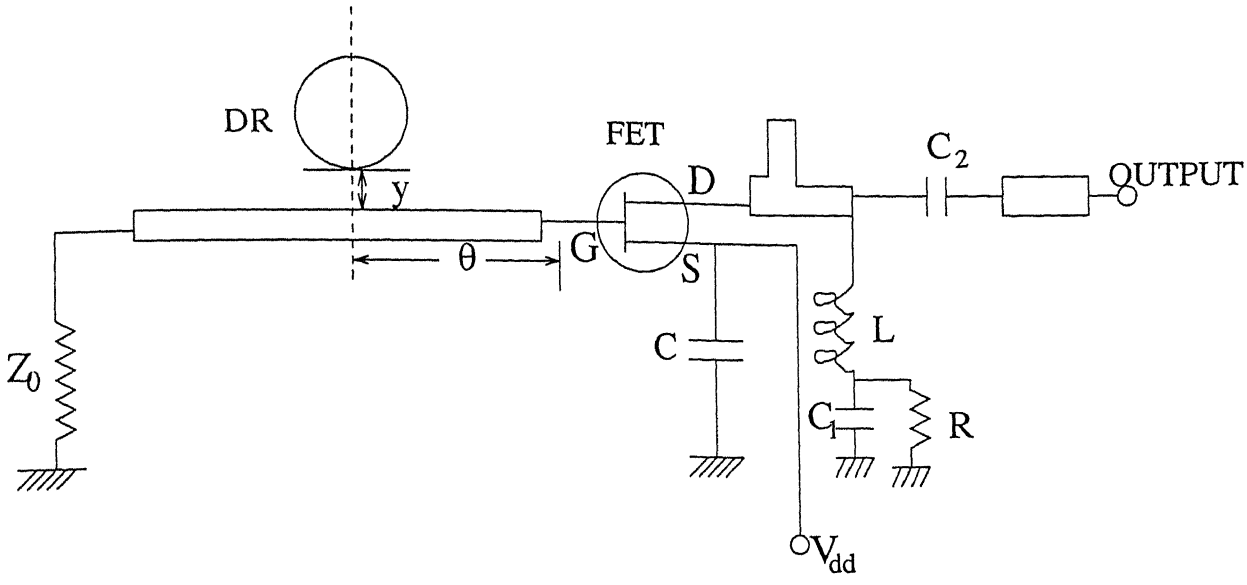

Figure 4.5: circuit diagram of oscillator [18]

R can be calculated as follow

$$\text{i=30 mA , } V_{dd} = 6V \text{ and } V_{ds} = 3V$$

$$\text{so } V_s\text{=3.0V and R= } V_s/\text{i or R=100}\Omega$$

4.4 Stability Consideration

The stability factor K decides whether the device will work as oscillator or not. From the S-parameter K is calculated as 0.3618. The value K is less than unity and device is potentially stable. However K is not much smaller than unity, and this mean that the choice of input and output impedance that will make the device oscillate is limited. In order to make the transistor more liable to oscillate at resonant frequency a capacitive feedback from source is added to ground, this capacitor also work as bypass capacitor.

This capacitor is normally chosen to have a reactance of the same order of magnitude as the input impedance of FET. For transistor NE70083, their are two source

of C is chosen the required microstrip stub can be easily calculated.

4.4.1 Calculation of S-parameter with feedback capacitor

First we convert device S-parameter to Z-parameter as follow

$$Z_m = Z_0(I + S_m)(I - S_m)^{-1} \tag{4.9}$$

Where Z_0 is the characteristic impedance of microstrip line and equal 50Ω. Converted Z-parameters are given below.

$$Z_m = \begin{bmatrix} 58.02\angle\text{-}73 & 4.25\angle\text{-}40 \\ 320.6\angle 27 & 9.78\angle\text{-}49 \end{bmatrix}$$

Since the input impedance of the device is of the order of Z_{11}=58.5$\angle -73^0$, The impedance of the feedback circuit Z_c (impedance of capacitance) = -j30Ω Then the feedback network can be represented by Z-matrix, Z_f, given by

$$Z_f = \begin{bmatrix} -j30 & -j30 \\ -j30 & -j30 \end{bmatrix}$$

The overall Z-matrix of transistor with this feedback capacitor is given by displayed equation:

$$Z = Z_m + Z_f$$

Or

$$Z = \begin{bmatrix} 87.34\angle\text{-}78.8 & 32.89\angle\text{-}84.3 \\ 308.1\angle 22 & 115.1\angle\text{-}58.8 \end{bmatrix}$$

Now the S-matrix corresponding to this Z-matrix can be obtained as follow.

$$S = (Z + Z_0 I)^{-1}(Z - Z_0 I) \tag{4.10}$$

The S-parameter of the transistor with feedback are

4.4.2 Design of open circuit stub to obtain capacitance at source terminals

X_c was chosen in the previous section as -j30Ω

$$Z_{oc}=-jZ_0\tan\beta l$$

Let $Z_0=60\Omega$

Here two parallel stub are designed at the source terminal of transistor. So X_c for each stub is

$$X_c=-j60\Omega$$

$$-j60=-j60\cot\beta l, \text{ or } \beta\, l=\pi/4$$

Assume microstrip line is a quasi TEM line, so

$$\beta=2\pi/\lambda, \text{ or } l=\lambda/8$$

$$\lambda_{(z_0=60)}= 5.7847 \text{ cm, so}$$

length of each stub (l_5)=7.23 mm

Width of stub (W)= 1.9 mm

4.5 Input and Output Matching Network

The equivalent circuit of input/output matching network is shown in figure 4.6. Using Y-parameter converted from S-parameter, the device impedance Z_D seen from terminal G can be calculated as [17].

$$Z_D = \frac{1 + Y_{22}Z_{out}}{Y_{11} + (Y_{11}Y_{22} - Y_{12}Y_{21})Z_{out}} \tag{4.11}$$

And the oscillation condition is given by

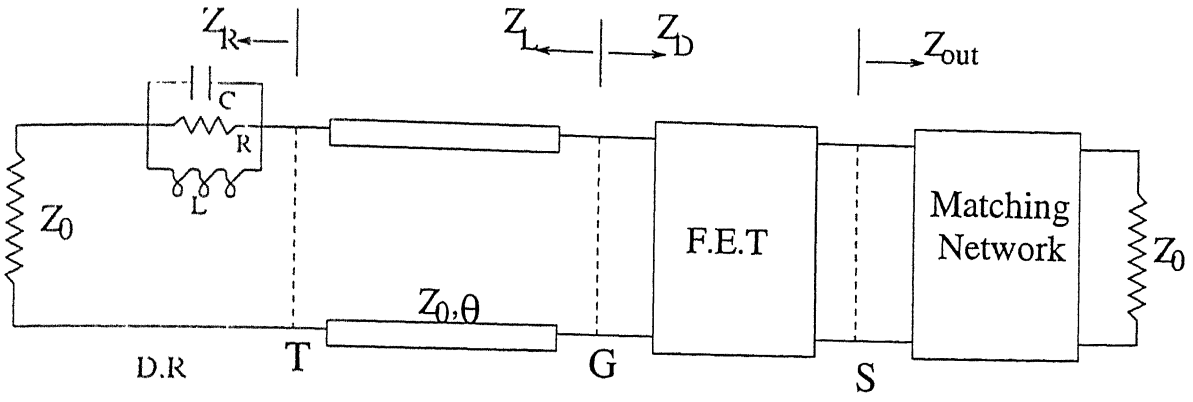

Figure 4.6: Equivalent Circuit of Oscillator [17]

From the two equations given above, the combination of output impedance, Z_{out}, and the device impedance, Z_D, is optimized for the maximum negative resistance of Z_D. The optimum combination of Z_{out} and Z_D is obtained as follows:

$$Z_{out}=5+j60(\Omega) \qquad Z_D=-97.8-j15.13(\Omega)$$

Once Z_{out} and Z_D are found output and input matching network can be designed.

4.5.1 Design of Output Matching Network

The output matching network is designed with a wider microstrip or lesser characteristic impedance ($Z_0 = 40\Omega$) line. Stub is designed with a high impedance line ($Z_0 = 60\Omega$). To minimize the transition interaction between the shunt stub and series transmission line, the shunt stub are usually balanced along the series transmission line. Two parallel shunt stubs of same admittance instead of the single stub must be provided for that purpose. The dimensions of stub and line are given below.

Length of the line (l_2)=1.058 cm Length of the each stub (l_3)=1.944 cm

4.5.2 Design of Input Matching Network

W(mm)	H(mm)	ϵ_{eff}	Z_0 (Ω)	$\lambda_g = \lambda_0/\sqrt{\epsilon_{eff}}$
1.9	0.8	1.83	60.0	5.7847
2.4	0.8	1.854	50.0	5.7468
4.0	0.8	1.91	40.0	5.6621

Table 4.1: Data for width and effective dielectric constant of microstrip

Z_R found from the smith chart is given by 102.5Ω.

The distance of DR from the Gate(Θ)=βl_1=0.097π radian so $l_1 = 0.485\lambda$

One end of the line is terminated with the matched load (50Ω), so that at any distance, this line offers pure resistance (50Ω). Distance of DR from the termination of line ((l_4) is chosen to be equal to distance of DR from the gate l_1. This makes the DR symmetric with the line as characterized in chapter 3 (figure 1.3) and effect of the side wall is also minimized. Now the distance between line and DR 'y' is selected in such a way that it simulates an impedance of 52.5Ω. From the from figure 3.13 this distance is found to be y=5.6mm. The PCB layout of the circuit is shown in figure 4.7.

Width of microstrip line (W), effective dielectric constant (ϵ_{eff}) and guided wavelength (λ_g) for various characteristic impedance (Z_0) are given in table4.4. Z_0 and ϵ_{eff} has been calculated using synthesis formula given in [17] and [19] respectively.

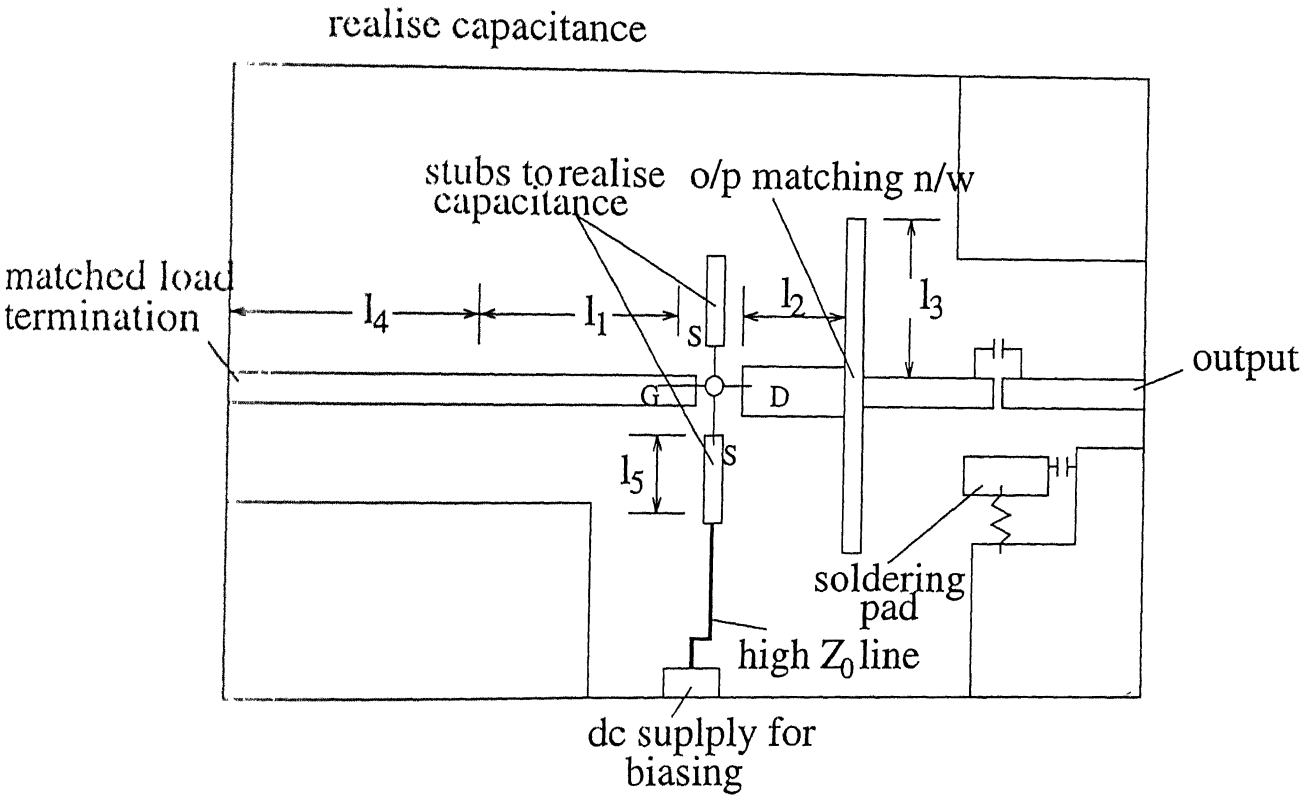

Figure 4.7: PCB layout of the oscillator circuit

Chapter 5

Conclusion and suggestion for future Development

5.1 Conclusion

In the design of microwave circuits, Q-factor is always a important parameter. Dielectric loss in the dielectric resonator (DR) and substrate and conductor loss in the top and bottom ground plate for the DR placed in the suspended substrate environment have been calculated in the present work. Thereafter, various Q-factor (Q_u, Q_c and Q_d) have been calculated. Subsequently, frequency dependent dielectric losses in DR and substrate have been incorporated. Effect of various parameters of DR and substrate on quality factor has also been studied.

An experimental procedure to measure the coupling between DR and microstrip for $TE_{01\delta}$ mode has been presented. Coupling (β), external quality factor (Q_{ext}), loaded quality factor (Q_L), Q_u and resonant frequency of DR in MIC environment have been measured. From these measured quantities, the equivalent circuit parameters of DR have been calculated.

Design of a oscillator containing DR as a frequency determining passive element has

Software has been developed to calculate the Q-factor and various losses. This software can be used in the design of microwave circuit containing the DR such as filters, mixer and oscillator.

5.2 Suggestion for future development

Following suggestion can be incorporated in any future work.

1. Experimental procedure to measure the coupling between DR and suspended substrate stripline for higher order mode can be developed.

2. Oscillator at higher frequency using DR can be designed.

Appendix A

Integral Constant:

The general expressions of I_1 and I_2 are

$$I_1 = \int_0^d r\ J_1^2(k_r r)\ dr \tag{A.1}$$

$$I_2 = \int_d^\infty r K_1^2(k_a r)\ dr \tag{A.2}$$

Where J_1 and K_1 are the Bassels function of first kind zero order and first kind first order respectively. The final expressions of I_1 and I_2 after integration are given by;

$$I_1 = \frac{d^2}{2}[J_1^2(k_r d) - J_0(k_r d) J_2(k_r d)] \tag{A.3}$$

$$I_2 = -\frac{J_0(k_r d)}{K_0(k_a d)} \frac{d^2}{2}[K_1^2(k_a d) - K_0(k_a d) K_2(k_a d)] \tag{A.5}$$

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